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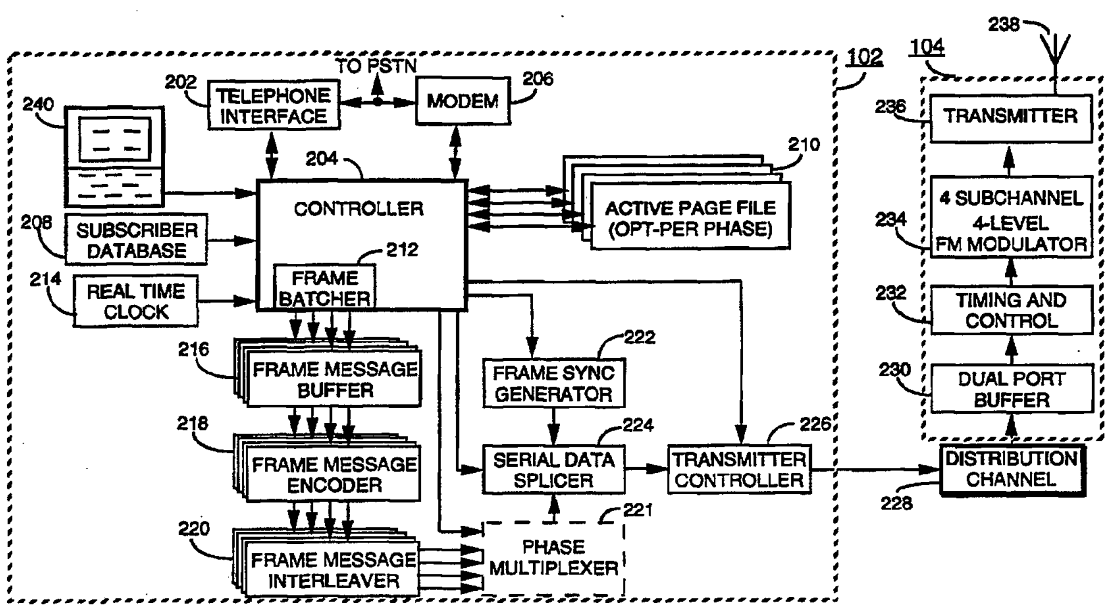

(57) Abstract

A communication system (100) broadcasting over a plurality of subchannels comprises a resource controller unit (204) having at least one of the plurality of subchannels serving as a control channel for addressing subscribers and directing them to receive messages or data on a set or a subset of the plurality of the subchannels, input means (240) for sending messages to the resource controller unit, and a selective call receiver (106) addressable by the resource controller unit, capable of receiving messages as directed by the resource controller on any of the subchannels and time slots directed by the resource controller.

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MULTIPLE SUBCHANNEL FLEXIBLE PROTOCOL METHOD AND APPARATUS

Technical Field

This invention relates generally to the field of communication systems and protocols and in particular, to a communication system and protocol that reallocates resources within multiple subchannels.

Background

There are many data communications systems in operation today which provide message distribution to data communication receivers, such as pagers. Many of these systems utilize signaling protocols which utilize time slots, or transmission frames, to which the pagers are assigned, thereby providing battery saving functions and other efficiencies during the normal course of message transmission. A paging terminal was provided in such systems to encode the received messages for transmission to the intended pagers during the assigned transmission frames. In signaling protocols such as the POCSAG signaling protocol, each time slot, or transmission frame, allowed for the transmission of only two code words, either both address code words, an address and a message code word, or two message code words. Since the transmission of even a simple telephone number required as a minimum two message code words, numeric message transmissions required on the average one and one-half frames, which periodically resulted in the inability to transmit address code words during the assigned transmission frames, because the transmission frame was filled by message code words associated with address code words transmitted in the previous transmission frame.

The above problem was alleviated in some signaling protocols by increasing the number of code words which could be transmitted in any transmission frame. However, when the number of transmittable code words is selected for the transmission frames, such transmission frames often have to be filled with idle code words when an insufficient number of messages have been received for transmission during any particular transmission frame. Such signaling protocols were also limited in the number of data communication receivers, or pagers, which could be assigned, or operated on any given channel, before the channel reached its maximum capacity. By increasing the transmission speed, additional pagers could be added to the system, however at the expense of a significant amount of unused channel capacity, until the system again filled up. Then, other systems resolved this problem (as described in pending applications assigned to the present assignee hereof and incorporated by reference herein by Schwendeman et al. having the Application No. 07/891,363 entitled "Data Communication Receiver Having Variable Length Message Carry-On" and by Kuznicki et al having the Application No. 07/891,503 and entitled "Data Communication Terminal Providing Variable Length Message Carry-On") by providing for a flexible system which enabled reconfiguring the amount of information which can be transmitted on the channel within the available transmission frames in order to maximize message throughput on a channel. These applications describe a flexible system which enables reconfiguring the amount of information which can be transmitted on a single channel within the available transmission frames. As the demand for greater capacity and throughput increases, there still exists a need for systems that make full use of subchannels to provide even greater throughput and flexibility than disclosed by the Applications by Kuznicki et al and Schwendeman et al. referred to herein above.

Summary of the Invention

A communication system broadcasting over a plurality of subchannels comprises a resource controller unit having at least

one of the plurality of subchannels serving as a control channel for addressing subscribers and directing them to receive messages or data on a set or a subset of the plurality of the subchannels, input means for sending messages to the resource controller unit, and a selective call receiver addressable by the resource controller unit, capable of receiving messages as directed by the resource controller on any of the subchannels and time slots directed by the resource controller.

In another aspect of the present invention, a method for receiving and decoding selective call messages transmitted in the form interleaved blocks of time divided signals on a plurality of subchannels to a plurality of selective call receivers, comprises the steps at one of the selective call receivers of decoding at least a first received block of information containing address and vector information for at least a first addressed message, at least a portion of the first received block being a control channel. Then, determining where the first addressed message will begin and the length of the first message from the address and vector information. And finally, decoding subsequent blocks of information on the plurality of subchannels to decode the first addressed message, the first addressed message being capable of residing in contiguous sections of blocks and portions of blocks on the plurality of subchannels.

Brief Description of the Drawings

FIG. 1 is an electrical block diagram of a data transmission system in accordance with the present invention.

FIG. 2 is an electrical block diagram of a terminal for processing and transmitting message information in accordance with the present invention.

FIGS. 3-5 are timing diagrams illustrating the transmission format of the signaling protocol utilized in accordance with the present invention.

FIGS. 6 and 7 are timing diagrams illustrating the synchronization signals utilized in accordance with the present invention.

FIG. 8 is an electrical block diagram of a data communication receiver in accordance with the present invention.

FIG. 9 is a more detailed electrical block diagram of the data communication receiver of FIG. 8 in accordance with the present invention.

FIG. 10 is an electrical block diagram of an alternative embodiment of a data communication receiver in accordance with the present invention.

FIG. 11 is a more detailed electrical block diagram of the data communication receiver of FIG. 10 in accordance with the present invention.

FIG. 12 is an electrical block diagram of yet another alternative embodiment of a data communication receiver in accordance with the present invention.

FIGs. 13-17 are diagrams illustrating the message capabilities of a system in accordance with the present invention.

Detailed Description of the Preferred Embodiment

FIG. 1 is an electrical block diagram of a data transmission system 100, such as a paging system, in accordance with the preferred embodiment of the present invention In such a data transmission system 100, messages originating either from a phone, as in a system providing numeric data transmission, or from a message entry device, such as an alphanumeric data terminal, are routed through the public switched telephone network (PSTN) to a paging terminal 102 which processes the numeric or alphanumeric message information for transmission by one or more transmitters 104 provided within the system. When multiple transmitters are utilized, the transmitters 104 preferably simulcast transmit the message information to data communication receivers 106. Processing of the numeric and alphanumeric information by the paging terminal 102, and the protocol utilized for the transmission of the messages is described below.

FIG. 2 is an electrical block diagram of the paging terminal 102 utilized for processing and controlling the transmission of the message information in accordance with the present invention. Short messages, such as tone-only and numeric messages which can be readily entered using a Touch-Tone telephone are coupled to the paging terminal 102 through a telephone interface 202 in a manner well known in the art. Longer messages, such as alphanumeric messages which require the use of a data entry device are coupled to the paging terminal 102 through a modem 206 using any of a number of well known modem transmission protocols. When a call to place a message is received, a controller 204 handles the processing of the message. The controller 204 is preferably a microcomputer, such as an MC68000 or equivalent, which is manufactured by Motorola Inc., and which runs various pre-programmed routines for controlling such terminal operations as voice prompts to direct the caller to enter the message, or the handshaking protocol to enable reception of messages from a data entry device. When a call is received, the controller 204 references information stored in the subscriber database 208 to determine how the message being received is to be processed. The subscriber data base 208 includes, but is not limited to such information as addresses assigned to the data communication receiver, message type associated with the address, and information related to the status of the data communication receiver, such as active or inactive for failure to pay the bill. A data entry terminal 240 is provided which couples to the controller 204, and which is used for such purposes as entry, updating and deleting of information stored in the subscriber data base 208, for monitoring system performance, and for obtaining such information as billing information.

The subscriber database 208 also includes such information as to what transmission frame and to what transmission phase the data communication receiver is assigned, as will be described in further detail below. The received message is stored in an active page file 210 which stores the messages in queue. Alternatively, the queue is provided in the active page file 210. The active page file 210 is preferably a dual

port, first in first out random access memory, although it will be appreciated that other random access memory devices, such as hard disk drives, can be utilized as well. Periodically the message information stored in each of the queue is recovered from the active page file 210 under control of controller 204 using timing information such as provided by a real time clock 214, or other suitable timing source. The recovered message information from the queue is sorted by frame number and is then organized by address, message information, and any other information required for transmission, and then batched into frames based upon message size by frame batching controller 212. The batched frame information is coupled to frame message buffers 216 which temporarily store the batched frame information until a time for further processing and transmission. Frames are batched in numeric sequence, so that while a current frame is being transmitted, the next frame to be transmitted is in the frame message buffer 216, and the next frame thereafter is being retrieved and batched. At the appropriate time, the batched frame information stored in the frame message buffer 216 is transferred to the frame encoder 218. The frame encoder 218 encodes the address and message information into address and message code words required for transmission, as will be described below. The encoded address and message code words are ordered into blocks and then coupled to a block interleaver 220 which interleaves preferably eight code words at a time for transmission in a manner well known in the art. The interleaved code words from each block interleaver 220 are then serially transferred on a bit by bit basis into a serial data stream by transmission phase. Optionally, if multiple phases are used, then the interleaved code words form each block interleaver 220 are then serially transferred to a phase multiplexer 221 (shown in ghost lines), which multiplexes the message information on a bit by bit basis into a serial data stream as before. The controller 204 next enables a frame sync generator 222 which generates the synchronization code which is transmitted at the start of each frame transmission. The synchronization code is multiplexed with address and message information under the control of controller

204 by serial data splicer 224, and generates therefrom a message stream which is properly formatted for transmission. The message stream is next coupled to a transmitter controller 226, which under the control of controller 204 transmits the message stream over a distribution channel 228. The distribution channel 228 may be any of a number of well known distribution channel types, such as wire line, an RF or microwave distribution channel, or a satellite distribution link. The distributed message stream is transferred to one or more transmitter stations 104, depending upon the size of the communication system. The message stream is first transferred into a dual port buffer 230 which temporarily stores the message stream prior to transmission. At an appropriate time determined by timing and control circuit 232, the message stream is recovered from the dual port buffer 230 and coupled to the input of preferably a 4 subchannel, 4-level FSK modulator 234. The modulated message stream is then coupled to the transmitter 236 for transmission via antenna 238.

FIGS. 3, 4 and 5 are timing diagrams illustrating the transmission format of the signaling protocol utilized in accordance with the preferred embodiment of the present invention. As shown in FIG. 3, the signaling protocol enables message transmission to data communication receivers, such as pagers, assigned to one or more of 128 frames which are labeled frame 0 through frame 127. It then will be appreciated that the actual number of frames provided within the signaling protocol can be greater or less than described above. The greater the number of frames utilized, the greater the battery life that may be provided to the data communication receivers operating within the system. The fewer the number of frames utilized, the more often messages can be queued and delivered to the data communication receivers assigned to any particular frame, thereby reducing the latency, or time required to deliver messages.

As shown in FIG. 4, the frames comprise a synchronization code (sync) followed preferably by eleven blocks of message information which are labeled block 0 through block 10. As

shown in FIG. 5. each block of message information comprises preferably eight address, control or data code words which are labeled word 0 through word 31 for each phase. Consequently, each phase in a frame allows the transmission of up to thirty-two (32) address, control and data code words. (In the case of 4 subchannel addressing, each phase in a frame allows the transmission of up to 4 times 32 words or 128 address, control and data code words. The address, control and data code words are preferably 31,21 BCH code words with an added thirty-second even parity bit which provides an extra bit of distance to the code word set. It will be appreciated that other code words, such as a 23,12 Golay code word could be utilized as well. Unlike the well known POCSAG signaling protocol which provides address and data code words which utilize the first code word bit to define the code word type, as either address or data, no such distinction is provided for the address and data code words in the signaling protocol utilized with the preferred embodiment of the present invention. Rather, address and data code words are defined by their position within the individual frames.

FIGS. 6 and 7 are timing diagrams illustrating the synchronization code utilized in accordance with the present invention. In particular, as shown in FIG. 6, the synchronization code comprises preferably three parts, a first synchronization code (sync 1), a frame information code word (frame info) and a second synchronization code (sync 2). As shown in FIG. 7, the first synchronization code comprises first and third portions, labeled bit sync 1 and BS1, which are alternating 1,0 bit patterns which provides bit synchronization, and second and fourth portions, labeled "A" and its complement "A bar", which provide frame synchronization. The second and fourth portions are preferably single 32,21 BCH code words which are predefined to provide high code word correlation reliability, and which are also used to indicate the data bit rate at which addresses and messages are transmitted. The table below defines the data bit rates which are used in conjunction with the signaling protocol.

Bit Rate	"A" Value
1600 bps	A1 and A1 bar
3200 bps	A2 and A2 bar
6400 bps	A3 and A3 bar
Not defined	A4 and A4 bar

As shown in the table above, three data bit rates are predefined for address and message transmission, although it will be appreciated that more or less data bit rates can be predefined as well, depending upon the system requirements. A fourth "A" value is also predefined for future use.

The frame information code word is preferably a single 32,21 BCH code word which includes within the data portion a predetermined number of bits reserved to identify the frame number, such as 7 bits encoded to define frame number 0 to frame number 127.

The structure of the second synchronization code is preferably similar to that of the first synchronization code described above. However, unlike the first synchronization code which is preferably transmitted at a fixed data symbol rate, such as 1600 bps (bits per second), the second synchronization code is transmitted at the data symbol rate at which the address and messages are to be transmitted in any given frame. Consequently, the second synchronization code allows the data communication receiver to obtain "fine" bit and frame synchronization at the frame transmission data bit rate.

In summary the signaling protocol utilized in accordance with an embodiment of the present invention comprises 128 frames which include a predetermined synchronization code followed by eleven data blocks which comprise eight address, control or message code words per phase. The synchronization code enables identification of the data transmission rate, and insures synchronization by the data communication receiver with the data code words transmitted at the various transmission rates.

The protocols described in the applications by Kuznicki et al and Schwendeman et al. are becoming known in the paging industry as the FLEX protocol. FLEX allows a communication

system to address and vector messages within a single channel, whereas the present invention allows a communication system to address and vector messages to one of N other subchannels in one embodiment, or in another embodiment, the communication system allows for the addressing and vectoring of messages to up to N subchannels simultaneously, where N can almost be any integer number . The following examples, for simplicity, illustrate embodiments where N=4, but of course, the scope of the claimed invention contemplates the embodiment where N can be any integer. For future reference, an embodiment where one of four subchannels can be addressed and vectored shall be called a 1X4 system, protocol, or receiver and an embodiment where four of four subchannels can be addressed and vectored simultaneously shall be called a 4X4 system, protocol, or receiver.

FIG. 8 is a block diagram of an embodiment of data communication receiver 106 in accordance with the present invention. The receiver 106 comprises of an antenna 802 coupled to a receiver module 804 which is coupled to a controller 816 via a 1X4 Decoder module 895 and a via synthesizer 899. The receiver 106 further includes memory 890 and input and output devices (885 & 880) as known in the art.

FIG. 9 is a more detailed electrical block diagram of the data communication receiver 106 shown in FIG. 8 in accordance with the present invention. The heart of the data communication receiver 106 is a controller 816, which is preferably implemented using an MC68HC11 microcomputer, such as manufactured by Motorola, Inc. The microcomputer controller, hereinafter called the controller 816, receives and processes inputs from a number of peripheral circuits, as shown in FIG. 9, and controls the operation and interaction of the peripheral circuits using software subroutines. The use of a microcomputer controller for processing and control functions is well known to one of ordinary skill in the art.

The data communication receiver 106 is capable of receiving address, control and message information, hereafter called "data" which is modulated using preferably 2-level and 4-

level frequency modulation techniques. The transmitted data is intercepted by an antenna 802 which couples to the input of a receiver section 804. Receiver section 804 processes the received data in a manner well known in the art, providing at the output an analog 4-level recovered data signal, hereafter called a recovered data signal. The recovered data signal is coupled to one input of a threshold level extraction circuit 808, and to an input of a 4-level decoder 810. The threshold level extraction preferably comprises two clocked level detector circuits (not shown) which have as inputs the recovered data signal. A level detector could detect the peak signal amplitude value and provide a high peak threshold signal which is proportional to the detected peak signal amplitude value, while another level detector detects the valley signal amplitude value and provides a valley threshold signal which is proportional to the detected valley signal amplitude value of the recovered data signal. Resistors are then utilized to enable decoding the 4-level data signals as will be described below.

When power is initially applied to the receiver portion, as when the data communication receiver is first turned on, a clock rate is preset to select a 128X clock, i.e. a clock having a frequency equivalent to 128 times the slowest data bit rate, which as described above is 1600 bps. The 128X clock is generated by 128X clock generator 844, as shown in FIG. 8, which is preferably a crystal controlled oscillator operating at 204.8 KHz (kilohertz). The output of the 128X clock generator 844 couples to an input of frequency divider 846 which divides the output frequency by two to generate a 64X clock at 102.4 KHz. The 128X clock allows the level detectors in the threshold level extraction circuit 808 to asynchronously detect in a very short period of time the peak and valley signal amplitude values, and to therefore generate the low (Lo), average (Avg) and high (Hi) threshold output signal values required for modulation decoding. After symbol synchronization is achieved with the synchronization signal, the controller 816 generates a second control signal to enable selection of a 1X symbol clock which is generated by symbol synchronizer 812 as shown in FIG. 9.

The most significant bit (MSB) output from the 4-level decoder 810 is coupled to an input of the symbol synchronizer 812 and provides a recovered data input generated by detecting the zero crossings in the 4-level recovered data signal. The symbol synchronizer 812 preferably uses the 64X clock at 102.4 KHz which is generated by frequency divider 846. A control signal (1600/3200) is provided to the symbol synchronizer 812 and is used to select the sample clock rate for symbol transmission rates of 1600 and 3200 symbols per second. The 1X and 2X symbol clocks are generated with 1600, 3200 and 6400 bits per second and are synchronized with the recovered data signal.

The 4-level binary converter 814 uses a 1X symbol clock and a 2X symbol clock along with the symbol output signals (MSB, LSB) and a selector signal (2L/4L) from the controller to select and provide control of the conversion of the symbol output signals as either 2-level FSK data, or 4-level FSK data. When the 2-level FSK data conversion (2L) is selected, only the MSB output is selected which is coupled to the input of a parallel to serial converter (not shown). When the 4-level FSK data conversion (4L) is selected, both the LSB and MSB outputs are selected which are coupled to the inputs of the parallel to serial converter.

Returning to FIG. 8, a serial binary data stream generated by the 4-level to binary converter 814 is coupled to inputs of a synchronization word correlator 818 and a demultiplexer 820. The synchronization word correlator has predetermined "A" word synchronization patterns that are recovered by the controller 816 from a code memory 822 and are coupled to an "A" word correlator (not shown). When the synchronization pattern received matches one of the predetermined "A" word synchronization patterns within an acceptable margin of error, an "A" or "A-bar" output is generated and is coupled to controller 816. The particular "A" or "A-bar" word synchronization pattern correlated provides frame synchronization to the start of the frame ID word, and also defines the data bit rate of the message to follow.

The serial binary data stream is also coupled to an input of the frame word decoder (not shown) which decodes the frame word and provides an indication of the frame number currently being received by the controller 816. During sync acquisition, such as following initial receiver turn-on, power is supplied to the receiver portion by battery saver circuit 848 which enables the reception of the "A" synchronization word, as described above, and which continues to be supplied to enable processing of the remainder of the synchronization code. The controller 816 compares the frame number currently being received with a list of assigned frame numbers stored in code memory 822. Should the currently received frame number differ from an assigned frame numbers, the controller 816 generates a battery saving signal which is coupled to an input of battery saver circuit 848, suspending the supply of power to the receiver portion. The supply of power will be suspended until the next frame assigned to the receiver, at which time a battery saver signal is generated by the controller 816 which is coupled to the battery saving circuit 848 to enable the supply of power to the receiver portion to enable reception of the assigned frame.

Returning to the operation of the synchronization correlator, a predetermined "C" word synchronization pattern is recovered by the controller 816 from a code memory 822 and is coupled to a "C" word correlator (not shown). When the synchronization pattern received matches the predetermined "C" word synchronization pattern with an acceptable margin of error, a "C" or "C-bar" output is generated which is coupled to controller 816. The particular "C" or "C-bar" synchronization word correlated provides "fine" frame synchronization to the start of the data portion of the frame. (See FIGs. 6 and 7).

The start of the actual data portion is established by the controller 816 generating a block start signal (Blk Start) which is coupled to inputs of a word de-interleaver 824.

Optionally, if multiple phases are used, then the block start signal is coupled to the inputs of a word de-interleaver 824 and a data recovery timing circuit 826. The block start signal is used to generate clocked phase signals which are synchronized with the

incoming message symbols. The clocked phase signal outputs of the phase timing generator 826 are coupled to inputs of a phase selector 828. During operation, the controller 816 recovers from the code memory 822, the transmission phase number to which the data communication receiver is assigned. The phase number is transferred to the phase select output (Ø Select) of the controller 816 and is coupled to an input of phase selector 828. A phase clock, corresponding to the transmission phase assigned, is provided at the output of the phase selector 828 and is coupled to clock inputs of the demultiplexer 820, block de-interleaver 824, and address and data decoders 830 and 832, respectively. The demultiplexer 820 is used to select the binary bits associated with the assigned transmission phase which are then coupled to the input of block de-interleaver 824, and clocked into the de-interleaver array on each corresponding phase clock.

The de-interleaver array is preferably a 32x32 bit array which de-interleaves thirty-two interleaved address, control or message code words, corresponding to one transmission block. The de-interleaved address code words are coupled to the input of address correlator 830. The controller 816 recovers the address patterns assigned to the data communication receiver, and couples the patterns to a second input of the address correlator. When any of the de-interleaved address code words matches any of the address patterns assigned to the data communication receiver within an acceptable margin of error, the message information associated with the address is then decoded by the data decoder 832 and stored in a message memory 850 in a manner well known to one of ordinary skill in the art. Following the storage of the message information, a sensible alert signal is generated by the controller 816. The sensible alert signal is preferably an audible alert signal, although it will be appreciated that other sensible alert signals, such as tactile alert signals, and visual alert signals can be generated as well. The audible alert signal is coupled by the controller 816 to an alert driver 834 which is used to drive an audible alerting device, such as a speaker or a transducer 836. The user can override the alert

signal generation through the use of user input controls 838 in a manner well known in the art.

Following the detection of an address associated with the data communication receiver, the message information is coupled to the input of data decoder 832 which decodes the encoded message information into preferably a BCD or ASCII format suitable for storage and subsequent display. The stored message information can be recalled by the user using the user input controls 838 whereupon the controller 816 recovers the message information from memory, and provides the message information to a display driver 840 for presentation on a display 842, such as an LCD display.

Referring to FIG. 10, another block diagram of an embodiment of the data communication receiver 106 in accordance with the present invention is shown. The receiver 106 comprises of an antenna 802 coupled to a receiver module 804 which is coupled to a controller 875 via a 4X4 Decoder module 897 and a via synthesizer 899. The receiver 106 further includes memory 890 and input and output devices (885 & 880) as known in the art. In one embodiment of the block diagram of FIG. 10, the receiver 106 would appear very much like the receiver of FIG. 9, except that the front end and decoder for the 4X4 FLEX receiver can appear as the block diagram shown in FIG. 11.

As before, the receiver 106 in FIG. 11 includes a receiver module 804 having an antenna 802. The receiver is coupled to a more sophisticated synthesizer 900 via a bank of mixers 310, 312, 314, and 316. The mixed signals from the bank of mixers is provided to the 4X4 decoder module 897. The module preferably comprises a bank of bandpass filters, detectors and decoders along with the appropriate amplification as is known in the art. Each bandpass filter (320, 322, 324 and 326 respectively) should be ideally designed to pass an appropriate subchannel on to their respective detectors (330, 332, 334, and 336) and their respective decoders (340, 342, 344, and 346). The signals from the decoder module 897 are then manipulated by the controller/data combiner in much the same manner as the controller 816 of FIG. 9. Of

course, the receiver 106 includes memory 890, and user input and output devices 885 and 880 respectively.

Referring to FIG. 12, another alternative embodiment is shown for the receiver 106 in FIG. 10 using a Digital Signal Processor (DSP) such as the Motorola DSP56001 or its functional equivalent. The receiver 106 of FIG. 12 preferably includes a linear receiver 404 having an antenna 402. Depending on the speed of the DSP, memory management and other housekeeping routines can be handled by the DSP. Otherwise, an optional controller 408, such as the controllers described in previous embodiments could be used. Ideally, in a 4X4 FLEX receiver, the DSP 406 will handle for four subchannels each: threshold level extraction, level synchronization, level synchronization correlation, data decoding, and data combining. Additionally, the DSP will also serve the functions of battery saving, de-multiplexing, de-interleaving, address correlation, phase selecting, and phase timing. Optionally, some of these tasks, and other tasks if needed, can be handled or shared by the controller 408. Finally, as usual, the receiver 106 includes memory 410, and user input and output devices 412 and 414 respectively.

FIGs. 13-17 illustrate typical timing diagrams associated with several embodiments in accordance with the present invention. FIG. 13 illustrates the timing diagram for a 1X4 FLEX receiver. The vector and addressing information will usually be found in the in the first subchannel, designated here as subchannel #0. Subchannel #0 or a portion of subchannel #0 will also be know as the control channel or the addressing channel. The vectoring information will usually designate what type of information will be received (suchas Hexidecimal or alphanumeric and whether the information will be provided on a single subchannel or on a multiple subchannel). The addressing information will designate what particular Word Number the message will start within the particular subchannel. In the case of a multiple subchannel message, the addressing information will designate what particular subchannel, block, and Word to begin providing the message. Additionally, the addressing can

designate corners in messages to provide further efficiency in messaging. Some of these features will become more apparent in the following discussions with regards to FIGS. 18-21.

Thus, in FIG. 13, the vectoring and addressing information in subchannel #0 directs the 1X4 Flex receiver to decode message #1 in subchannel 2 at a particular block and word. It should be understood that the 1X4 Flex receiver could have been directed to decode message #1 in any one of the available (four in this instance) subchannels, not just the subchannel shown. In FIG. 14, the vectoring and addressing information in subchannel #1 directs a 4X4 Flex receiver to decode the repeated message #1 in each of the subchannels at different blocks and words. Further note as illustrated in FIGs. 13 and 14, that it is within contemplation of the present invention that the control channel can reside on any of the subchannels (the intermediary subchannels and the highest or last subchannel), not just subchannel #0 (the lowest or first subchannel) as shown in FIG. 13.

In FIG. 15, the vectoring and addressing information in this case directs a Flex 4X4 receiver to decode three different sized messages (message #1, #2 and #3) at different subchannels. In the case of message #1, the message is decoded at a later time frame portion within subchannel #0. Message #2 is decoded in portions of contiguous "areas" within subchannels #1 and #2, while message #3 is decoded in portions of subchannel 3.

FIG. 16 is the same illustration as FIG. 15, but further illustrating the blocks and block boundaries preferably associated with the present invention. As shown, there are preferably 8 blocks within the block boundaries which are decoded at a time. The addressing and vectoring information in subchannel #0 first decoded by the 4X4 Flex receiver will direct the receiver to decode message #1 at the beginning at block 8 of subchannel #0, message #2 at the beginning of block 6 of subchannel #1 (and ending at block 10 of subchannel #2), and message #3 at the beginning of block 4 of subchannel 3.

Of course, the present invention maintains the flexibility found in the embodiments of Kuznicki et al and Schwendeman et

al in terms of being able to send and receive messages at variable speeds, but additionally, the present invention allows the receiver to send both 1X4 Flex messages and 4X4 Flex messages and further pack them within a 4 subchannel format that provides great efficiency as shown in FIG. 17. In this case, the receiver is shown as decoding messages in blocks of 5, where there are 11 blocks to a frame as previously described with regard to FIGs. 3-7. (Please note these are not the only formats available contemplated for use with the claimed invention.) When the receiver decodes the first five blocks (probably the last block of Frame N-1 and the first four blocks of Frame N), the receiver decodes the vectoring and addressing information first, found here in this case in the first block and portion of the second block of the second block of Frame N of subchannel #0. The vectoring information would indicate that the first four messages are 1X4 messages found in a single subchannel (subchannel #0), while the messages #5-#8 are 4X4 Flex messages decoded throughout portions of the four subchannels.

Operationally, a 1X4 FLEX receiver receiving message #4 (MSG4) as shown in FIG. 17, would decode block #0 and detect the message's address from the portion 2 of the block #0 and perhaps the message's vector from portion 3. Additionally, the receiver might decode block #1 and retrieve further vector information from portion 3. (Portion 1 of block #0 preferably contains Block Information Words). The 1X4 FLEX receiver would then decode blocks #2, #3 and #4 in sequence to extract its message.

Again referring to FIG. 17, a 4X4 FLEX receiver, receiving messages MSG5, MSG6, MSG7, and MSG8, would decode block #0 and block #1 to extract the address and vectors and then would decode the blocks from the start to end of the message based upon the vector information. This device would demodulate and decode data from multiple subchannels simultaneously as required based upon the subchannels used in transmission of the message.

A single device could decode all the messages in FIG. 17 in a variety of sized segments (for example, 5 blocks). The device

preferably has a receiver that decodes the addressing and vectoring information in blocks 1 and 2 in subchannel #0, then the message #3, along with portions of message #4 and portions of message #5, then message #1 along with more portions of message #4 and #5 and the beginning portions of message #6, then the remainder of message #2, with portions of message #4 and message #6 along with the remainder of message of #5. As the next segment of blocks are decoded, the remainder of messages #4 and #6 are decoded, and the entire message #7 is decoded. Additionally, a portion of message #8 is decoded. Finally, after the next segment of blocks are decoded, the remainder of message #8 is decoded by the receiver.

The 8 messages can be so efficiently packed together because of the flexibility of the protocol which allows for mixing of protocols and "corner" commands to make room for single subchannel messages (pages) among multiple subchannel messages or pages. These capabilities and advantages will become further apparent in the explanation of the format of the 32 (or more, actually 64 in the examples to follow) bit words used for vectoring and addressing the incoming messages at a receiver as shown in the tables below.

TABLE 1

HEX / Binary Vector (Single Subcarrier)

1 2 3 4 5 6 7 21	31	32
Information	Parity	Ck
x_0 x_1 x_2 x_3 V_0 V_1 V_2 V_3 b_0 b_1 b_2 b_3 b_4 b_5 b_6 b_7 b_8 y_0 y_1 m_0 s_0	p p p p p p p p p p	P
x_0 x_1 x_2 x_3 n_0 n_1 n_2 n_3 n_4 n_5 n_6 n_7 n_8 s_7 s_6 s_5 s_4 s_3 s_2 s_1 s_0	p p p p p p p p p p	P

V - Vector Type $V_3V_2V_1V_0$ = 0110 - HEX Vector Single Subcarrier
b - Word Number of message start $b_8b_7b_6b_5b_4b_3b_2b_1b_0$ (1 - 511 Decimal)
y - Subchannel assigned
m - 0 implies message is in this frame
1 implies message in future frame. s_7 - s_1 contains the frame num.
n - Number of message words $n_8n_7n_6n_5n_4n_3n_2n_1n_0$ (1 to 511 Decimal)
s - Spares
x - Std 4 bit Check Character

Table 1 shows the formatting for a HEX/Binary message using a 1X4 Flex format, which requires the designation of a vector type (HEX vector, single subchannel), the Word number where the message will start, the number of message words in the particular frame, and the subchannel assigned.

TABLE 2

HEX / Binary Vector (Multiple Subcarrier)

1 2 3 4 5 6 7 .21 Information	. . . . .31 Parity	32 Ck
$x_0\ x_1\ x_2\ x_3\ V_0\ V_1\ V_2\ V_3\ b_0\ b_1\ b_2\ b_3\ b_4\ b_5\ b_6\ b_7\ b_8\ b_9\ b_{10}\ d_0 d_1$	p p p p p p p p p p	P
$x_0\ x_1\ x_2\ x_3\ n_0\ n_1\ n_2\ n_3\ m_0\ c_0\ c_1\ c_2\ c_3\ c_4\ c_5\ e_0\ e_1\ e_2\ e_3\ e_4\ e_5$	p p p p p p p p p p	P

V - Vector Type $V_3V_2V_1V_0$ = 1110 - HEX Vector Multiple Subcarrier
b - Word of message start = $b_{10}b_9$-Subchannel, $b_8b_7b_6b_5$-Block, $b_4b_3b_2b_1b_0$-Word
d - Number of additional corners in message field (6 bits/corner)
n - Number of message words = $n_{10}n_9n_8n_7n_6n_5n_4n_3n_2n_1n_0$ (1 to 511 Decimal)
n_{10} - n_4 are in message field.
m - 0 implies message is in this frame
1 implies message is in future frame. c_0-e_0 represent the frame num.
The first corner information is in message field.
c - first Corner = c_5c_4-Subchannel, $c_3c_2c_1c_0$- Block
e - second Corner = e_5e_4-Subchannel, $e_3e_2e_1e_0$- Block
x - Std 4 bit Check Character

Table 2 illustrates the formatting for a HEX/Binary message using the 4X4 Flex format, which requires the designation of a vector type (HEX/Binary, multiple subchannel), the location where the first Word of the message will begin including information detailing the subchannel, block, and specific Word within the block where the message will begin, the number of message words, and cornering information which further defines within a frame and block of information where a particular message will reside when it is decoded by the receiver. The cornering information, as in the message start information, should include the subchannel location, block location, and in some cases the Word location. Note that there may be other more precise or more flexible methods within the scope and spirit of the present

invention than using "corner" commands , but "corner" commands are an efficient compromise

TABLE 3

Alphanumeric Vector (Single Subchannel)

1 2 3 4 5 6 7 21	31	32
Information	Parity	Ck
x_0 x_1 x_2 x_3 V_0 V_1 V_2 V_3 b_0 b_1 b_2 b_3 b_4 b_5 b_6 b_7 b_8 y_0 y_1 m_0 s_0	p p p p p p p p p p	P
x_0 x_1 x_2 x_3 n_0 n_1 n_2 n_3 n_4 n_5 n_6 n_7 n_8 s_7 s_6 s_5 s_4 s_3 s_2 s_1 s_0	p p p p p p p p p p	P

V - Vector Type $v_3v_2v_1v_0$ = 0101 - Alpha Vector Single Subcarrier
b - Word Number of message start $b_8b_7b_6b_5b_4b_3b_2b_1b_0$ (1 - 511 Decimal)
y - Subchannel assigned
m - 0 implies message is in this frame
1 implies message in future frame. s_7 - s_1 contains the frame number.
n - Num. of message words in this frame $n_8n_7n_6n_5n_4n_3n_2n_1n_0$ (1 to 511 Decimal)
s - Spares
x - Std 4 bit Check Character

Table 3 shows the formatting for a alphanumeric message using a 1X4 Flex format, which requires the designation of a vector type (alphanumeric, single subchannel), the Word number where the message will start, the number of message words in the particular frame, and the subchannel assigned.

TABLE 4

Alphanumeric Vector (Multiple Subchannel)

1 2 3 4 5 6 7 21	31	32
Information	Parity	Ck
x_0 x_1 x_2 x_3 V_0 V_1 V_2 V_3 b_0 b_1 b_2 b_3 b_4 b_5 b_6 b_7 b_8 b_9 b_{10} d_0 d_1	p p p p p p p p p p	P
x_0 x_1 x_2 x_3 n_0 n_1 n_2 n_3 m_0 c_0 c_1 c_2 c_3 c_4 c_5 e_0 e_1 e_2 e_3 e_4 e_5	p p p p p p p p p p	P

V - Vector Type $v_3v_2v_1v_0$ = 1101 - HEX Vector Multiple Subcarrier
b - Word of message start = $b_{10}b_9$-Subchannel, $b_8b_7b_6b_5$-Block, $b_4b_3b_2b_1b_0$- Word
d - Number of additional corners in message field (6 bits/corner)
n - Number of message words = $n_{10}n_9n_8n_7n_6n_5n_4n_3n_2n_1n_0$ (1 to 511 Decimal)
n_{10} - n_4 are in message field.
m - 0 implies message is in this frame
1 implies message is in future frame. c_0-e_0 represent the frame num.
The first corner information is in message field.
c - 1st Corner = c_5c_4-Subchannel, $c_3c_2c_1c_0$- Block
e - 2nd Corner = e_5e_4-Subchannel, $e_3e_2e_1e_0$- Block
x - Std 4 bit Check Character

Table 4 illustrates the formatting for an alphanumeric message using the 4X4 Flex format, which requires the designation of a vector type (alphanumeric, multiple subchannel), the location where the first Word of the message will begin including information detailing the subchannel, block, and specific Word within the block where the message will begin, the number of message words, and cornering information which as before, further defines within a frame and block of information where a particular message will reside when it is decoded by the receiver.

The present invention has been described in detail in connection with the disclosed embodiments. These embodiments, however, are merely examples and the invention is not restricted thereto. It will be understood by those skilled in the art that variations and modifications can be made within the scope and spirit of the present invention as defined by the appended claims.

What is claimed is:

Claims

1. A communication system broadcasting over a plurality of subchannels, comprising:

a resource controller unit having at least one of the plurality of subchannels serving as a control channel for addressing subscribers and directing them to receive messages or data on a set or a subset of the plurality of the subchannels;

input means for sending messages to the resource controller unit; and

a selective call receiver addressable by the resource controller unit, capable of receiving messages as directed by the resource controller on any of the subchannels and time slots directed by the resource controller.

2. The communication system of claim 1 wherein the control channel further addresses and directs subscribers to receive messages or data on a given time slot within a TDD frame and within a set or subset of the plurality of subchannels or within a portion of one of the plurality of subchannels.

3. A terminal in a communication system broadcasting over a plurality of subchannels, comprising:

a resource controller unit having at least one of the plurality of subchannels serving as a control channel for addressing subscribers and directing them to receive messages or data on a set or a subset of the plurality of the subchannels;

input means for sending messages to the resource controller unit; and

a transmitter for addressing a selective call receiver addressable by the resource controller unit, capable of receiving messages as directed by the resource controller on any of the subchannels and time slots directed by the resource controller.

4. The terminal of claim 3, wherein the control channel resides within a portion of the lowest or first subchannel and is used to address and vector messages to the remaining portion of the lowest subchannel and to the other subchannels.

5. A selective call receiver capable of receiving messages broadcast over a plurality of subchannels, comprising:

a receiver module capable of receiving selective call signals on the plurality of subchannels and providing a received signal;

a decoder module coupled to the receiver module;

a bank of mixers for mixing the received signal from the receiver module with an injection signal from a synthesizer to provide a mixed signal to the decoder module, the decoder module comprising of a corresponding bank of bandpass filters, detectors and decoders for each of the plurality of subchannels; and

a controller for controlling the decoder module, the synthesizer, a sensory alert device, and a display device all coupled to the controller.

6. A selective call receiver capable of receiving messages broadcast over a plurality of subchannels, comprising:

a receiver module capable of receiving selective call signals on the plurality of subchannels and providing a received signal;

a digital signal processor coupled to the receiver module for providing the function for each subchannel of threshold level extraction, level synchronization, level synchronization correlation, data decoding, and data combining and for providing the general functions of battery saving, de-multiplexing, de-interleaving, address correlation, phase selecting, and phase timing.

7. A method for queuing messages for transmission in a data communication terminal having an input for receiving messages and for assigning the same into a plurality of transmission frames and subchannels assigned for transmission to a plurality of data communication receivers, said method comprising the steps of:

storing the received messages in a first memory area;

generating periodic timing signals;

recovering the stored messages from the first memory in response to the periodic timing signals being generated;

queuing the recovered messages for the assigned transmission frame and subchannel into a second memory area having a predetermined queue capacity;

monitoring the second memory queue capacity, and expected queue capacities for one or more subsequent transmission frames and subchannels;

determining when the messages stored within the second memory area exceed the predetermined queue capacity;

storing the excess messages recovered in a third memory area;

generating designating information designating one or more subsequent transmission frames and subchannels during which the excess messages stored in the third memory area are to be transmitted; and

transmitting the messages and designating information stored in the second memory area within the assigned transmission frame and subchannel; and

transmitting the excess messages stored in the third memory area within the one or more subsequent transmission frames designated by the designating information.

8. A method for receiving and decoding selective call messages transmitted in the form of interleaved blocks of time divided signals on a plurality of subchannels to a plurality of selective call receivers, comprising the steps at one of the selective call receivers:

decoding at least a first received block of information containing address and vector information for at least a first addressed message, at least a portion of the first received block being a control channel;

determining where the first addressed message will begin and the length of the first message from the address and vector information; and

decoding subsequent blocks of information on the plurality of subchannels to decode the first addressed message, the first addressed message being capable of residing in contiguous sections of blocks and portions of blocks on the plurality of subchannels.

9. The method for receiving and decoding selective call messages of claim 8, wherein the step of decoding subsequent blocks of information further comprises the step of:

simultaneously demodulating and decoding information on the plurality of subchannels as directed by the address and vector information.

10. The method of claim 8, wherein the address and vector information contains start addresses and message lengths for a plurality of messages, wherein the plurality of messages can span a plurality of subchannels and a plurality of blocks within a time frame.

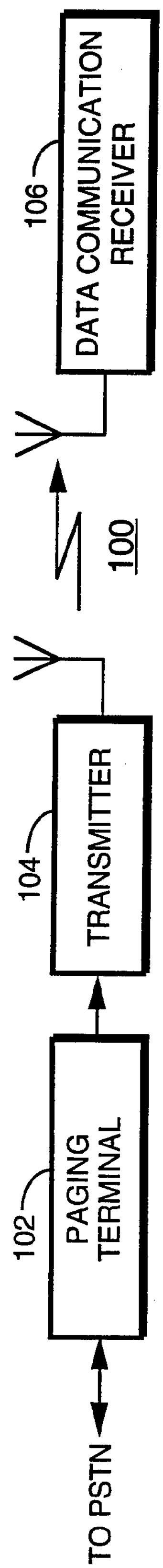
TO PSTN
PAGING TERMINAL
102
TRANSMITTER
104
DATA COMMUNICATION RECEIVER
106
100

FIG. 1

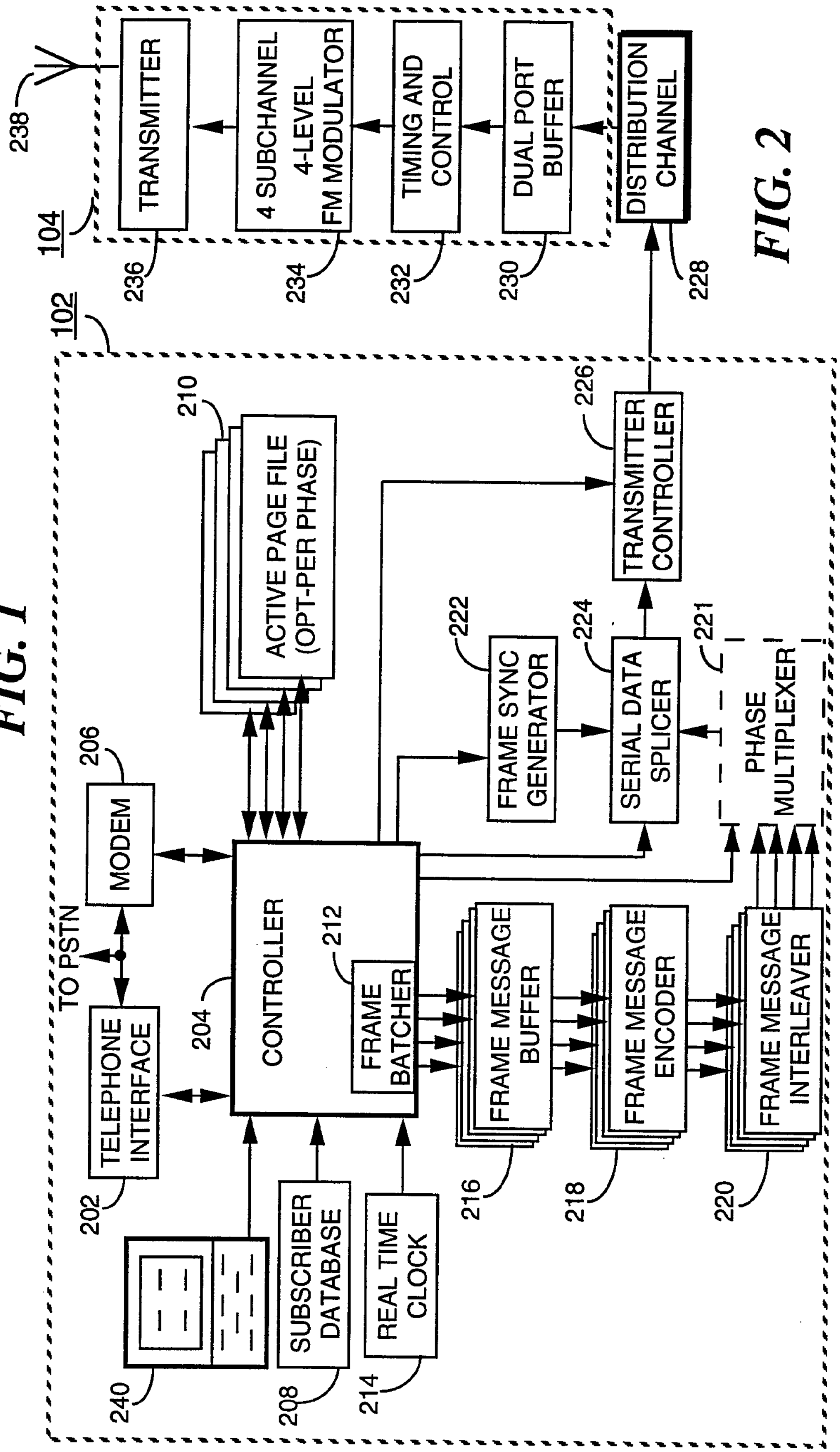
102
TO PSTN
TELEPHONE INTERFACE
202
MODEM
206
CONTROLLER
204
FRAME BATCHER
212
ACTIVE PAGE FILE (OPT-PER PHASE)
210
240
SUBSCRIBER DATABASE
208
REAL TIME CLOCK
214
FRAME MESSAGE BUFFER
216
FRAME MESSAGE ENCODER
218
FRAME MESSAGE INTERLEAVER
220
FRAME SYNC GENERATOR
222
SERIAL DATA SPLICER
224
PHASE MULTIPLEXER
221
TRANSMITTER CONTROLLER
226
DISTRIBUTION CHANNEL
228
104
TRANSMITTER
236
238
4 SUBCHANNEL 4-LEVEL FM MODULATOR
234
TIMING AND CONTROL
232
DUAL PORT BUFFER
230

FIG. 2

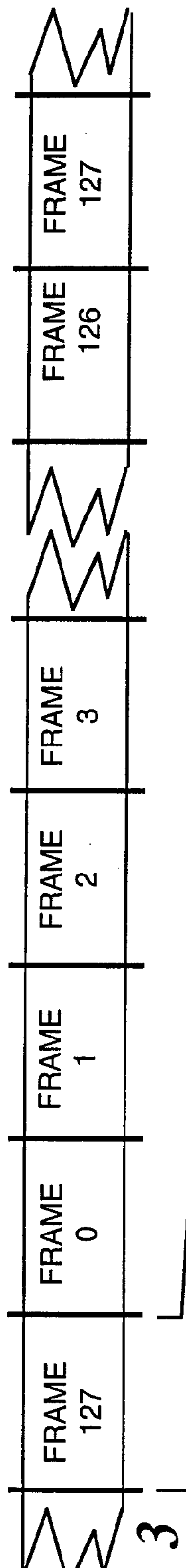

FIG. 3

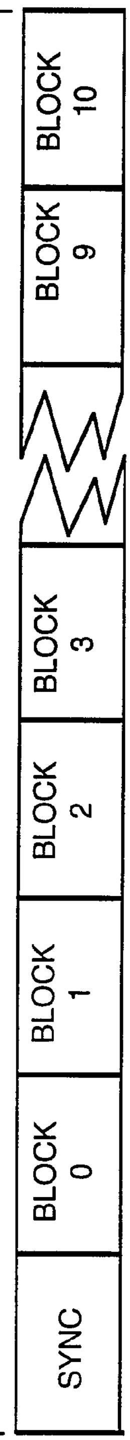

FIG. 4

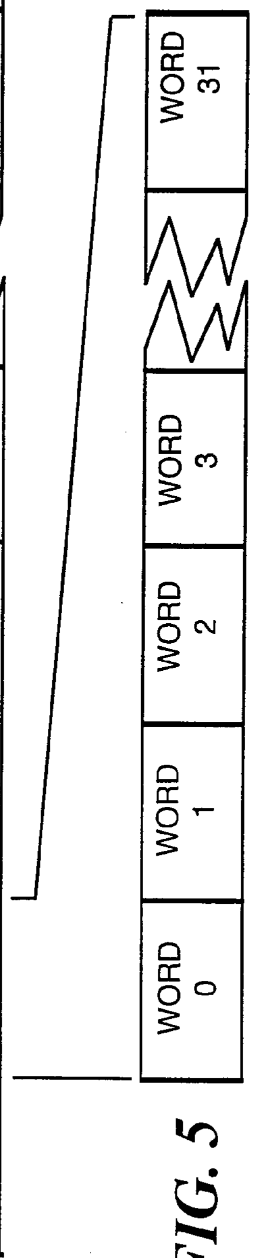

FIG. 5

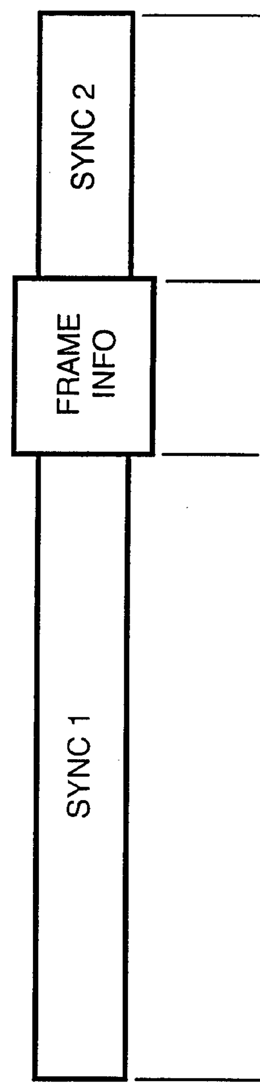

FIG. 6

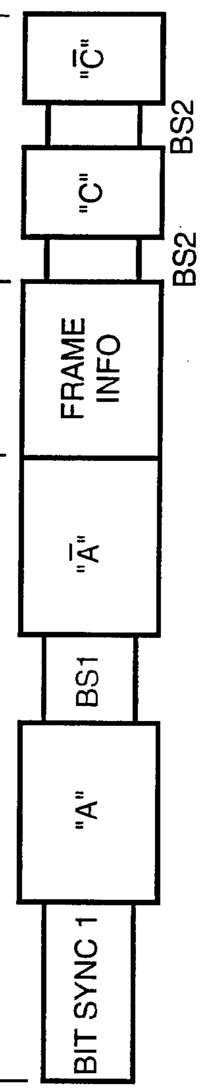

FIG. 7

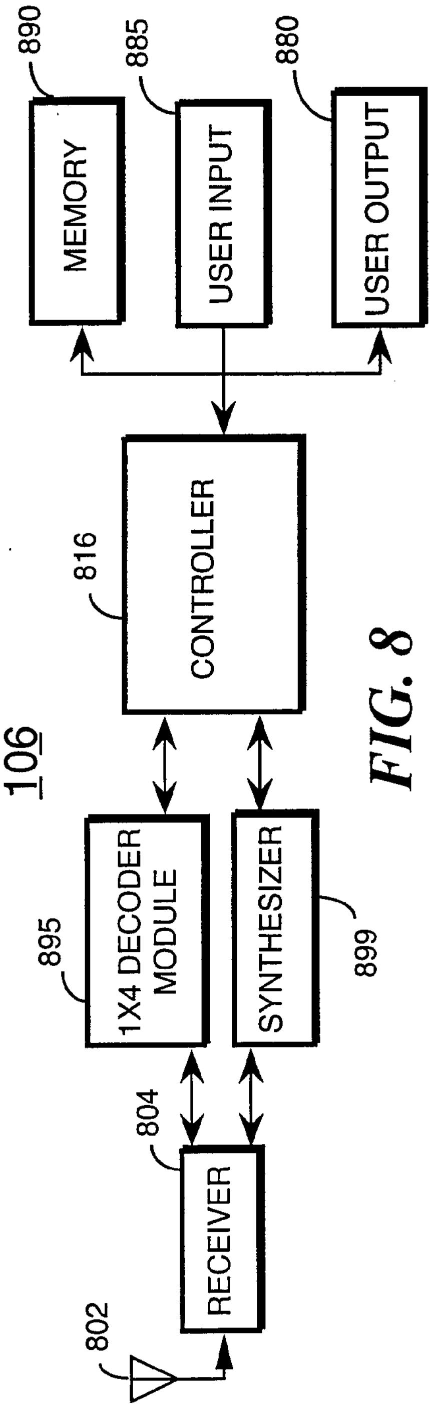
106
802
RECEIVER
804
1X4 DECODER MODULE
895
SYNTHESIZER
899
CONTROLLER
816
MEMORY
890
USER INPUT
885
USER OUTPUT
880

FIG. 8

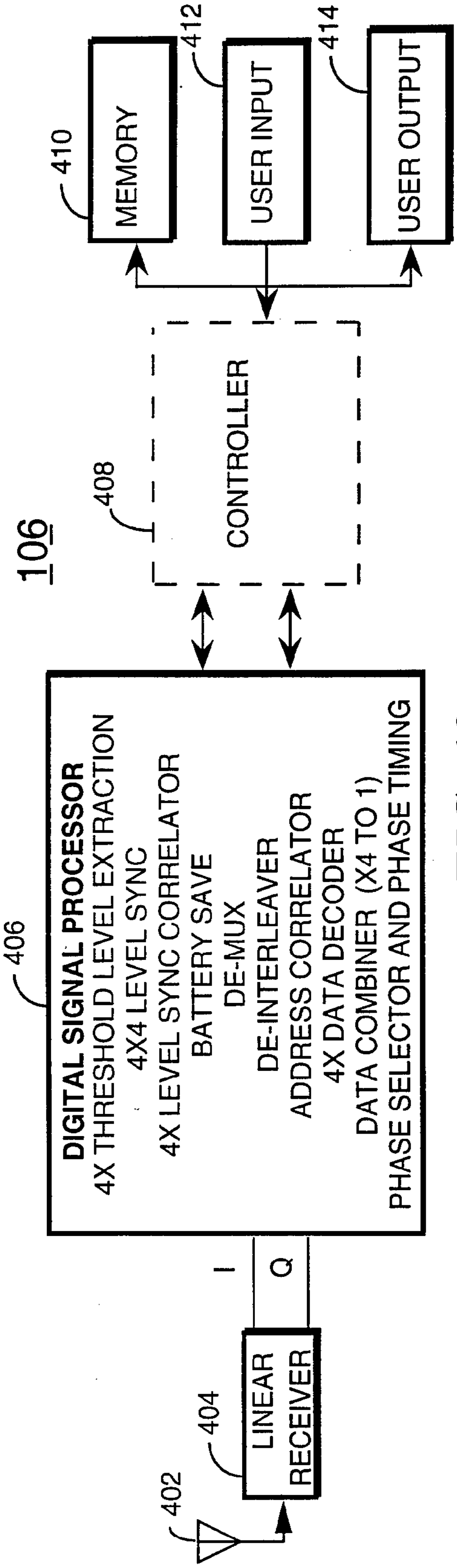
106
402
LINEAR RECEIVER
404
I
Q
DIGITAL SIGNAL PROCESSOR
4X THRESHOLD LEVEL EXTRACTION
4X4 LEVEL SYNC
4X LEVEL SYNC CORRELATOR
BATTERY SAVE
DE-MUX
DE-INTERLEAVER
ADDRESS CORRELATOR
4X DATA DECODER
DATA COMBINER (X4 TO 1)
PHASE SELECTOR AND PHASE TIMING
406
CONTROLLER
408
MEMORY
410
USER INPUT
412
USER OUTPUT
414

FIG. 12

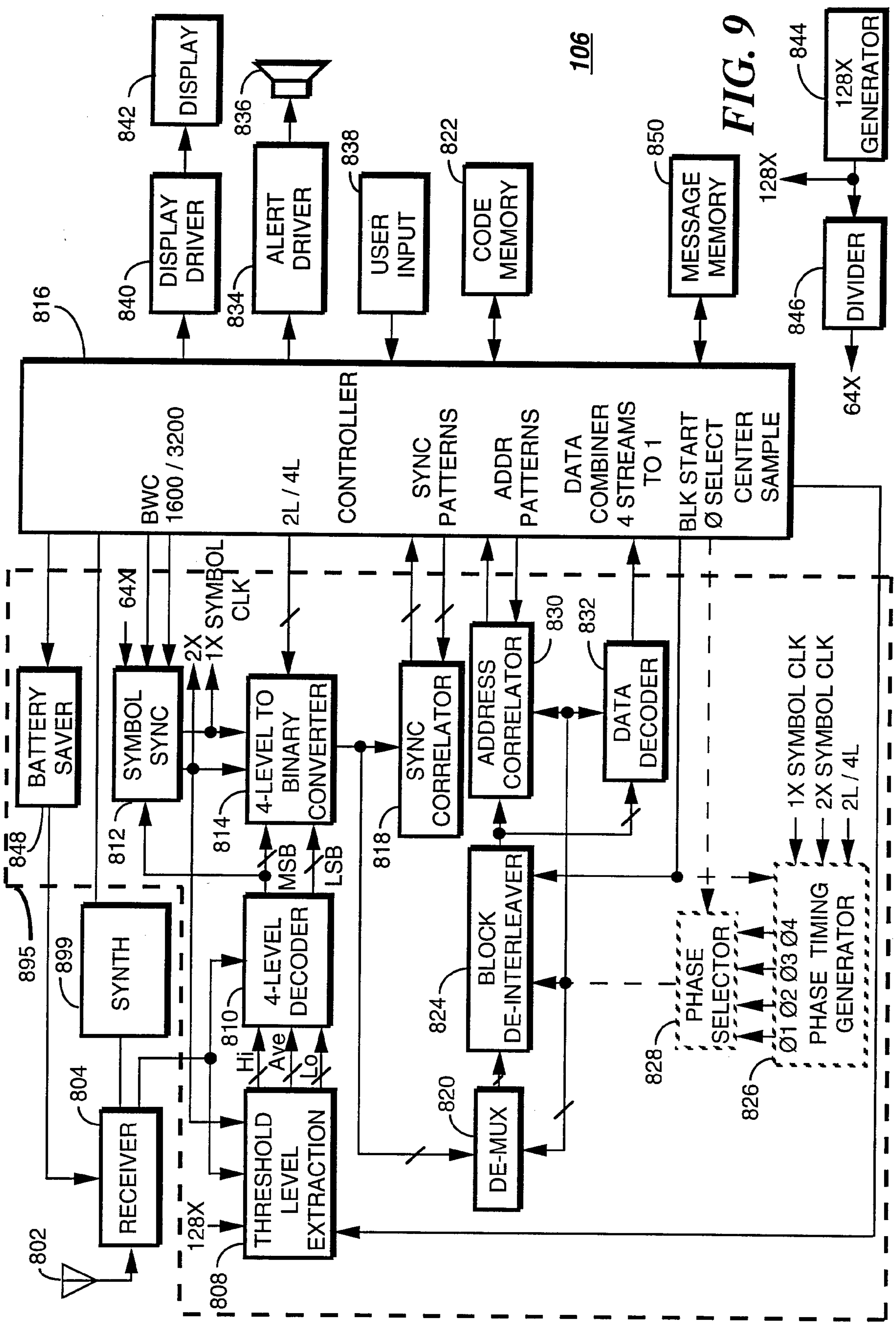
106
DISPLAY
842
DISPLAY DRIVER
840
ALERT DRIVER
834
836
USER INPUT
838
CODE MEMORY
822
MESSAGE MEMORY
850
128X GENERATOR
844
128X
DIVIDER
846
64X
816
CONTROLLER
BWC 1600 / 3200
2L / 4L
SYNC PATTERNS
ADDR PATTERNS
DATA COMBINER 4 STREAMS TO 1
BLK START
Ø SELECT
CENTER SAMPLE
BATTERY SAVER
848
SYMBOL SYNC
812
64X
2X
1X SYMBOL CLK
4-LEVEL TO BINARY CONVERTER
814
MSB
LSB
SYNC CORRELATOR
818
ADDRESS CORRELATOR
830
DATA DECODER
832
BLOCK DE-INTERLEAVER
824
DE-MUX
820
PHASE SELECTOR
828
Ø1 Ø2 Ø3 Ø4
PHASE TIMING GENERATOR
826
1X SYMBOL CLK
2X SYMBOL CLK
2L / 4L
895
SYNTH
899
4-LEVEL DECODER
810
Hi
Ave
Lo
RECEIVER
804
802
THRESHOLD LEVEL EXTRACTION
808
128X

FIG. 9

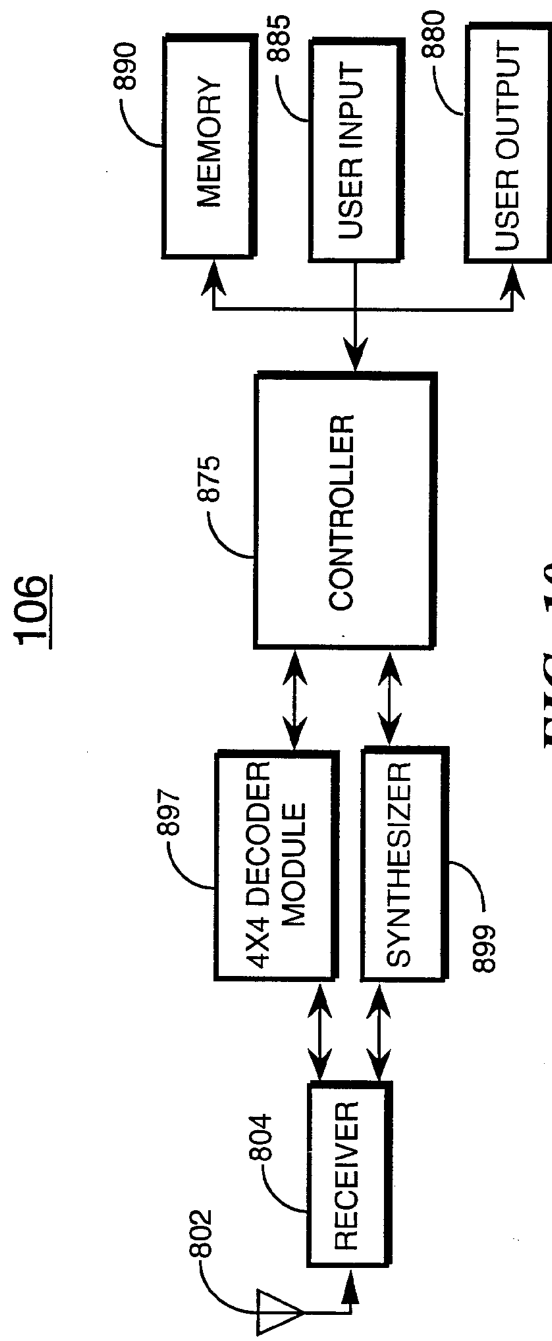
106
890
MEMORY
885
USER INPUT
880
USER OUTPUT
875
CONTROLLER
897
4X4 DECODER MODULE
SYNTHESIZER
899
804
RECEIVER
802

FIG. 10

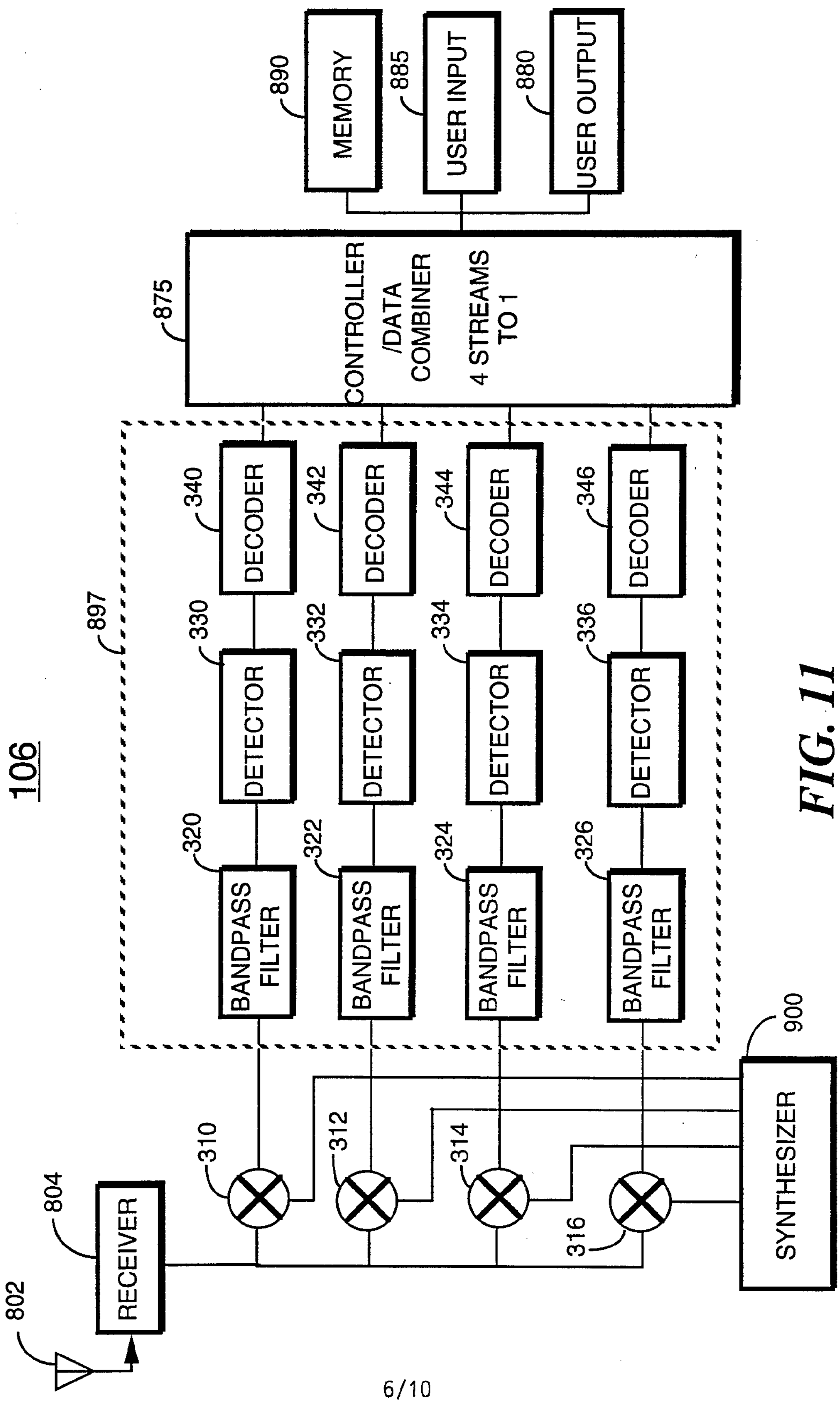
106
802
804
RECEIVER
310
312
314
316
SYNTHESIZER
900
897
BANDPASS FILTER
320
322
324
326
DETECTOR
330
332
334
336
DECODER
340
342
344
346
875
CONTROLLER /DATA COMBINER
4 STREAMS TO 1
890
MEMORY
885
USER INPUT
880
USER OUTPUT

FIG. 11

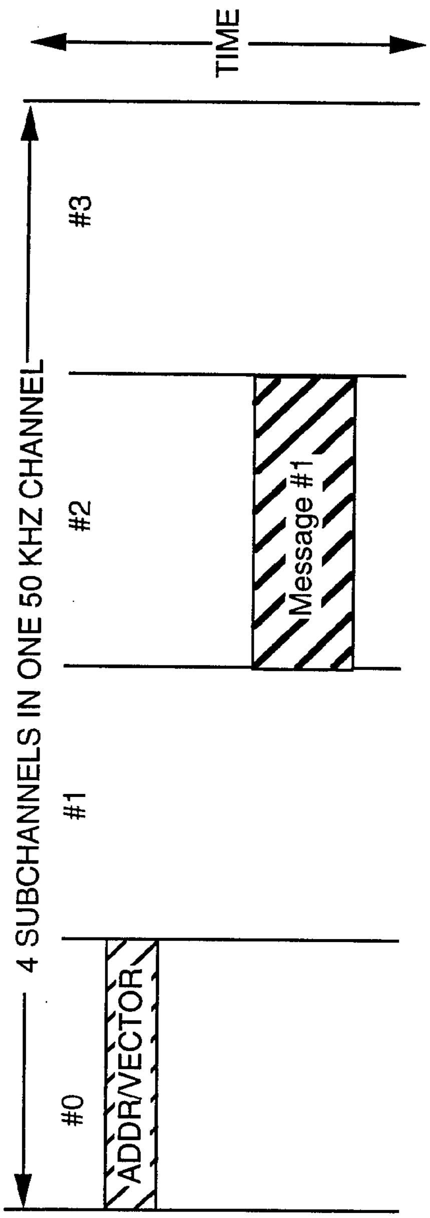
TIME
4 SUBCHANNELS IN ONE 50 KHZ CHANNEL
#0
#1
#2
#3
ADDR/VECTOR
Message #1

FIG. 13

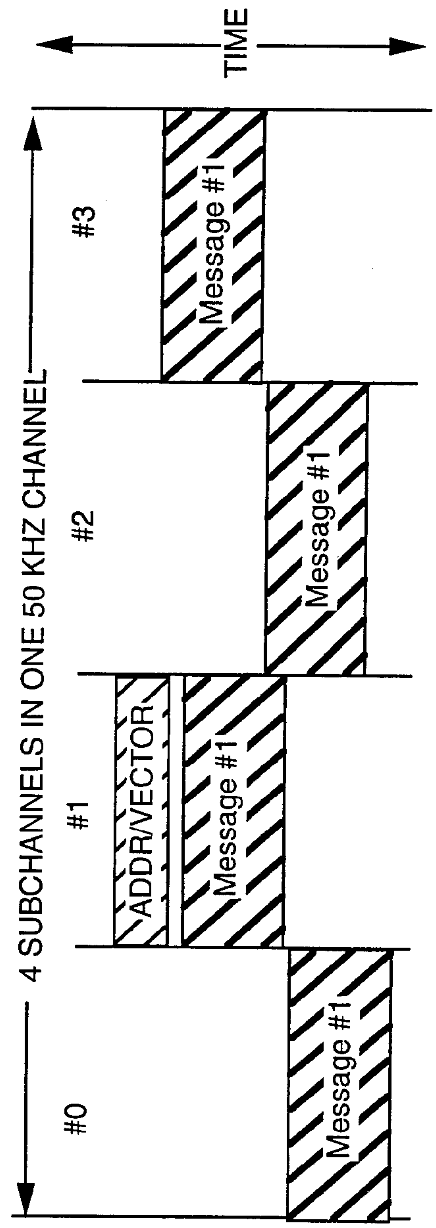
TIME
4 SUBCHANNELS IN ONE 50 KHZ CHANNEL
#0
#1
#2
#3
ADDR/VECTOR
Message #1
Message #1
Message #1
Message #1

FIG. 14

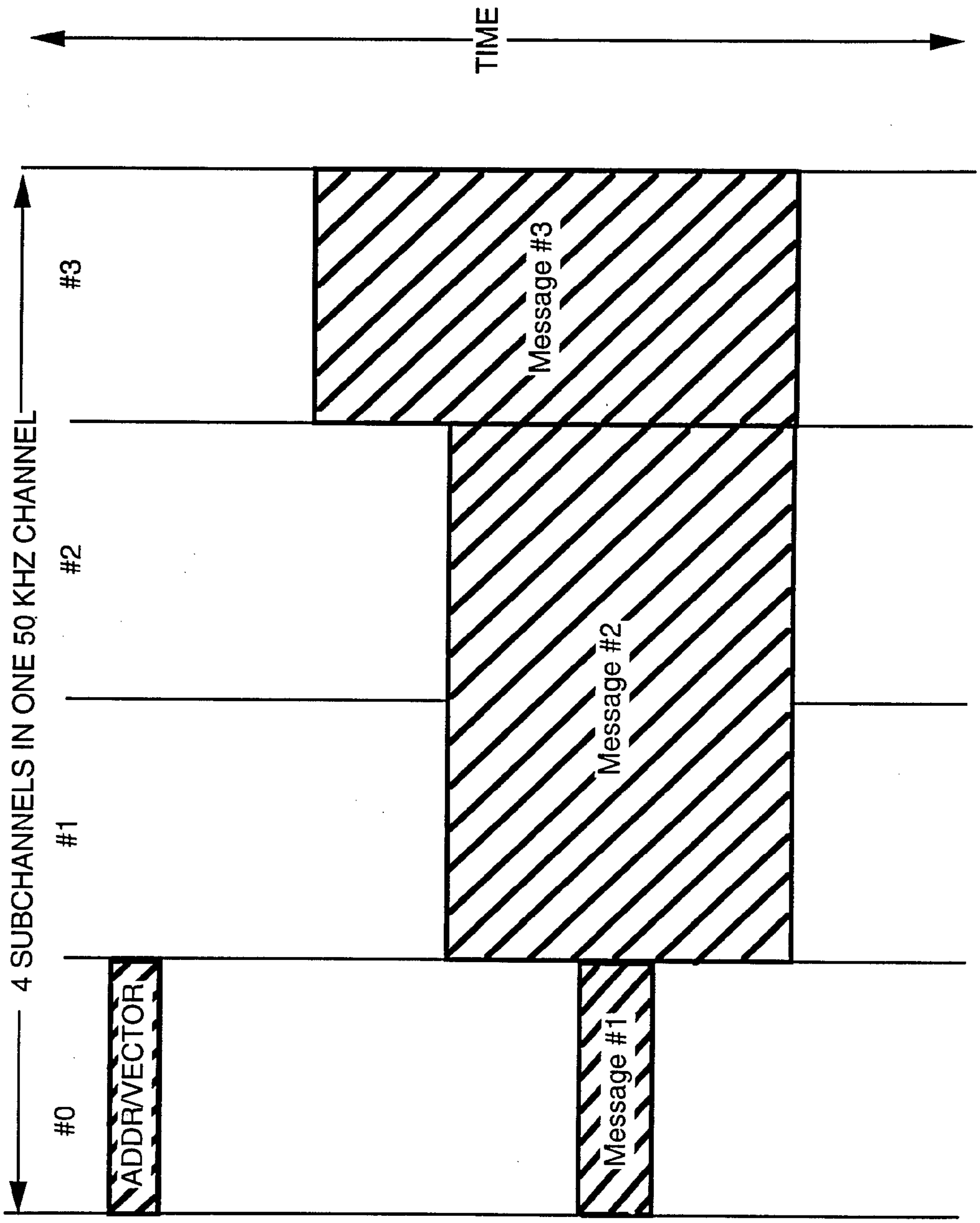
TIME
4 SUBCHANNELS IN ONE 50 KHZ CHANNEL
#0
#1
#2
#3
ADDR/VECTOR
Message #1
Message #2
Message #3

FIG. 15

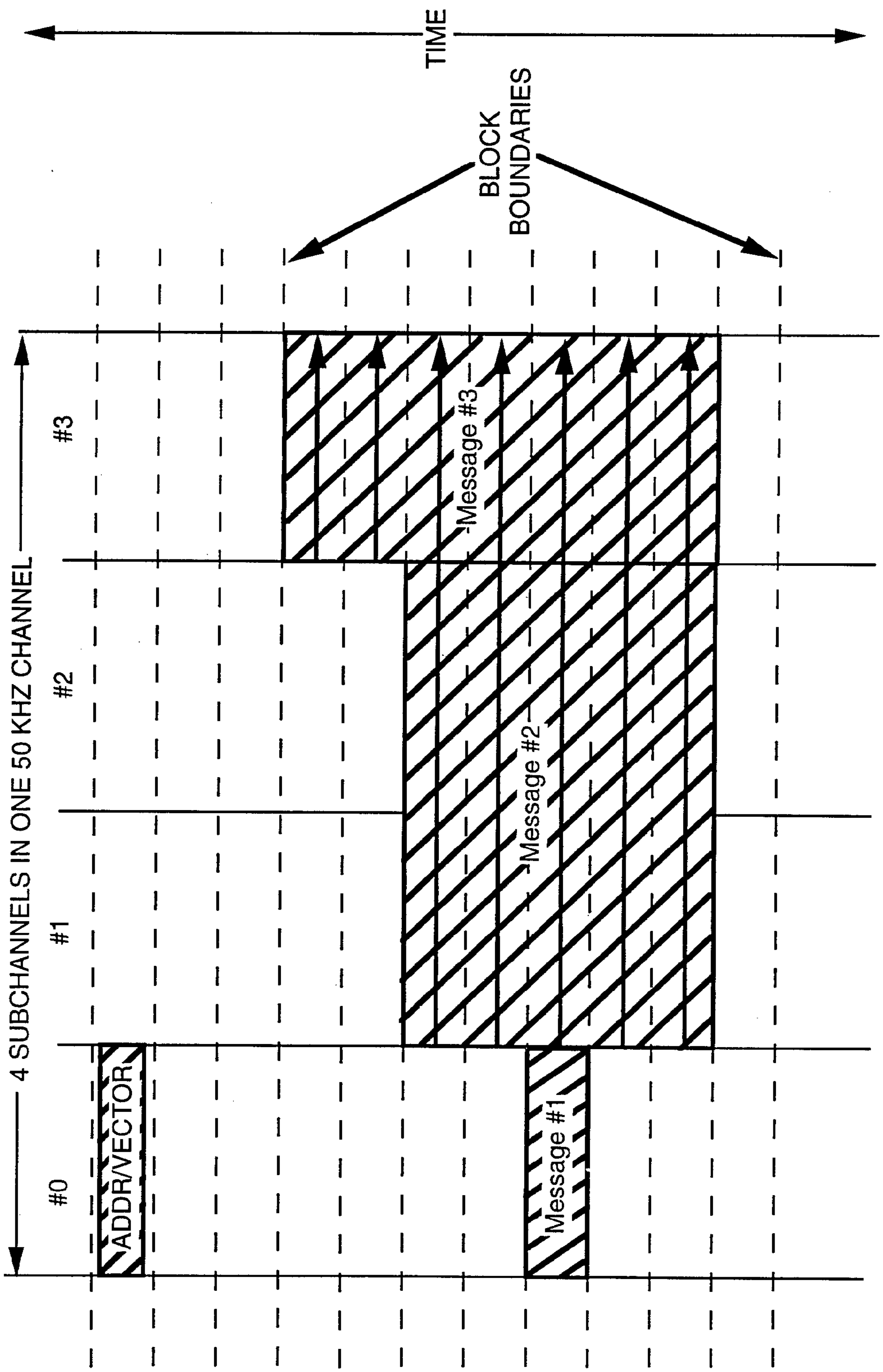
TIME
BLOCK BOUNDARIES
4 SUBCHANNELS IN ONE 50 KHZ CHANNEL
#0
#1
#2
#3
ADDR/VECTOR
Message #1
Message #2
Message #3

FIG. 16

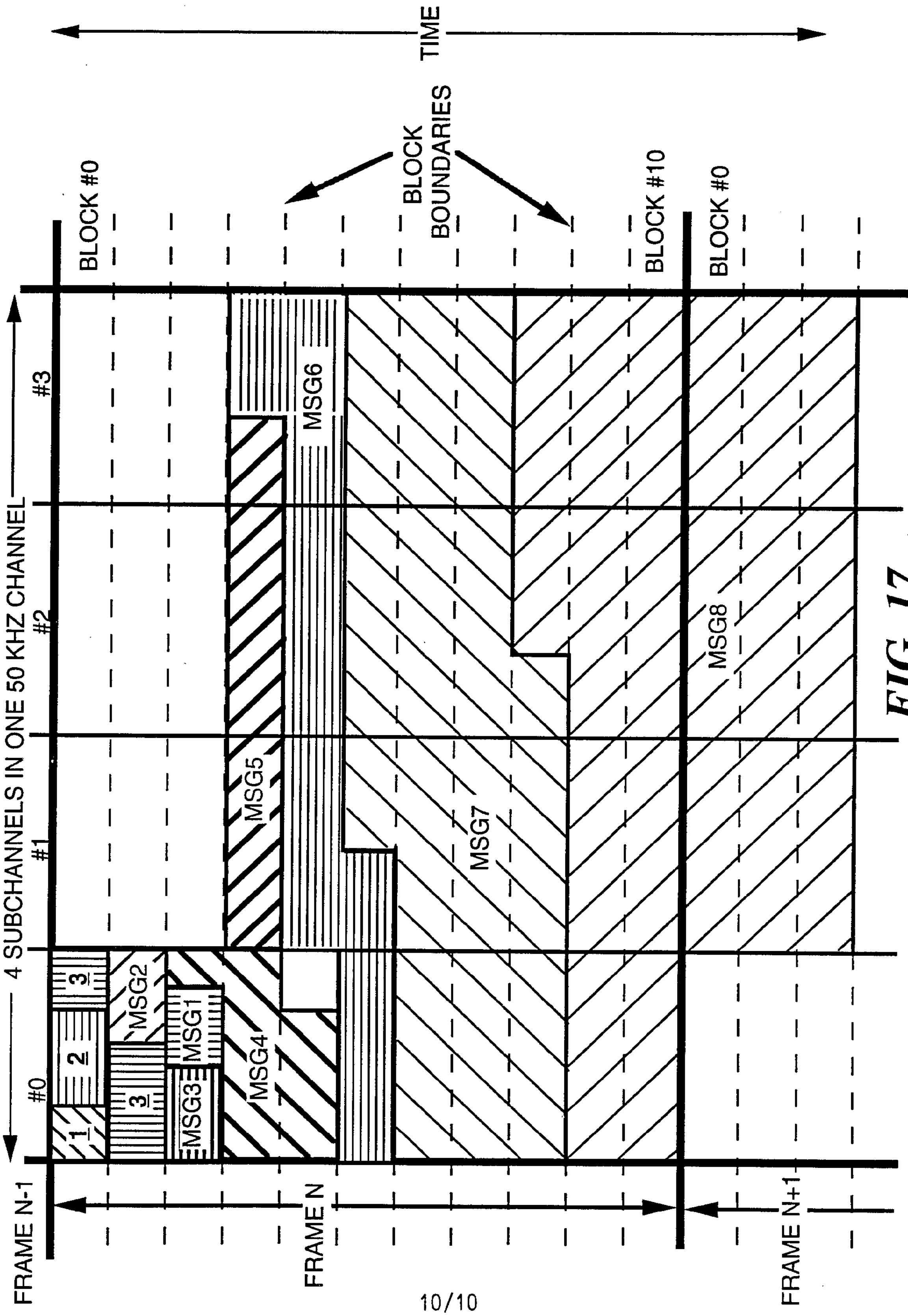
TIME
BLOCK BOUNDARIES
BLOCK #0
BLOCK #10
BLOCK #0
4 SUBCHANNELS IN ONE 50 KHZ CHANNEL
#0
#1
#2
#3
1
2
3
3
MSG1
MSG2
MSG3
MSG4
MSG5
MSG6
MSG7
MSG8
FRAME N-1
FRAME N
FRAME N+1

FIG. 17

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US95/05377

A. CLASSIFICATION OF SUBJECT MATTER

IPC(6) :H04Q 7/00
US CL :340/825.44; 379/57; 370/95.1
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 340/825.44; 379/57, 58, 60, 62; 370/94.1, 95.1

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
Y	US, A, 4,903,320 (HANAWA) 20 FEBRUARY 1990, COL. 1 LINES 27-46.	1-22
Y	US, A, 4,949,395 (RYDBECK) 14 AUGUST 1990, ABSTRACT, COL. 1 LINE 19 - COL. 2 LINE 19.	1-22
A	US, A, 5,278,890 (BEESON, JR. ET AL) 11 JANUARY 1994, ABSTRACT.	1
A	US, A, 5,260,944 (TOMABECHI) 09 NOVEMBER 1993, ABSTRACT.	1
A	US, A, 5,199,031 (DAHLIN) 30 MAY 1993, ABSTRACT.	1

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

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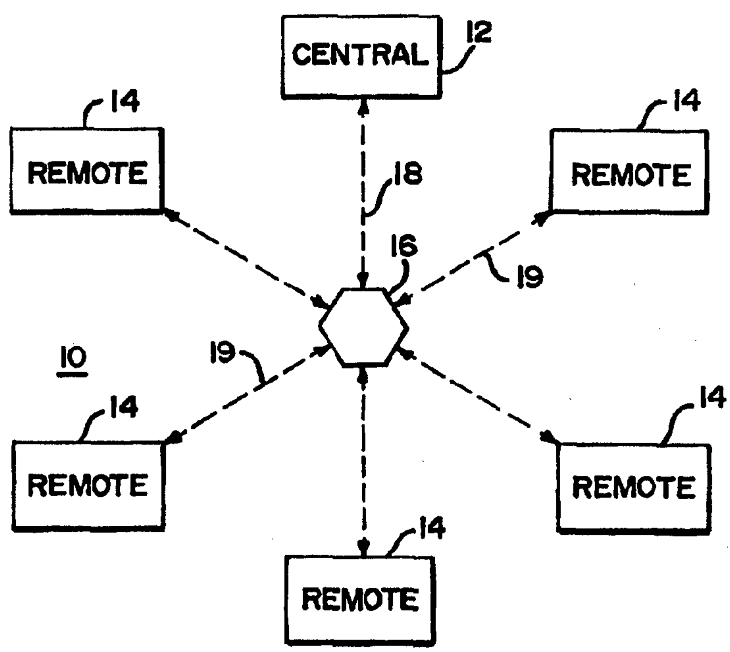

(57) Abstract

An apparatus is provided for exchanging data between a central station (12) and a plurality of wireless remote stations (14) on a time division multiple access communication channel. The apparatus includes means for receiving access requests from remote stations of the plurality of remote stations (14) during a first time interval under a contention based protocol and a non-contention based protocol and means for polling for data transfers during a second time period remote stations (14) of the plurality of remote stations providing access requests under non-contention based protocols during the first time period.

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APPARATUS FOR EXCHANGING DATA BETWEEN A CENTRAL STATION AND A PLURALITY OF WIRELESS REMOTE STATIONS

Background of the Invention

This invention generally relates to the field of data communications networks. More particularly, this invention pertains to a multiple access protocol for a data communications network having a number of users exchanging data between individual remote stations to a central station over a single optical infrared channel.

A multipoint digital communications network typically consists of a number of remote stations which communicate with a central station over one or more two-way communications channels. For example, personal computers are typically connected to a wide variety of peripherals or other computers via wire cables, i.e., a hard-wired communications link. Moreover, local area networks (LAN's) are often used to integrate remote terminals that are located at the same site. Depending upon the number of users, distance between terminals, number of peripherals, frequency of system reconfiguration, portability of the remote stations, etc., the hard-wired cable system may not be practical for a given application. Hence, various wireless communication technologies have been employed, particularly when

a system includes a large number of users and/or portable, hand-held computer devices.

Among the more common wireless technologies are narrow-band radio frequency (RF) systems, spread spectrum RF, ultrasonic, and infrared optical. Radio frequency systems are often significantly degraded by electromagnetic noise and interference, as well as by large signal amplitude variations and multipath interference. Moreover, RF systems are typically subject to governmental licensing and regulation. Alternative wireless systems employing ultrasonic sound waves experience severe problems with the complete loss of signals due to nulls in the transmission path.

Optical-infrared communications, however, is not affected by electromagnetic interference, and is much less susceptible to multipath interference. Furthermore, optical systems are inherently secure (since the infrared light does not penetrate walls), have no known health or safety effects, and are not subject to F.C.C. regulation. Moreover, infrared transceivers draw relatively low currents, which is particularly important with respect to hand-held battery-powered portable computers. Thus, the use of infrared light as the wireless medium is well suited to such applications.

In order for the remote stations to communicate with the central station, the remote stations must be able to gain access to the commonly-shared communications channel using some type of multiple-access signalling or control protocol. As used in the data communications field, a "protocol" is a formal set of rules governing the format and control of inputs and outputs

between two communicating devices in order to ensure the orderly transfer of information. Typical multiple-access protocols may be categorized into two broad classes: contention-based protocols (i.e., random access), and noncontention-based protocols, (i.e., scheduled access). Contention-based protocols are characterized in that any remote user with a data message can contend for the channel by transmitting its data message immediately in an on-demand fashion, taking the chance that no other remote stations will transmit at the same time and thus collide with it. When a collision occurs, the data message is seldom received correctly, if at all. Since there is no coordination between contending remote stations, the number of collisions dramatically increases as the number of users increase, or as the channel load increases. Hence, contention-based protocols are not suitable for many data communications applications.

Noncontention-based protocols are characterized in that they provide the necessary coordination between the remote stations to ensure that no two remote stations transmit at the same time to contend for the channel. In other words, the users in a noncontention system take turns accessing the network in an orderly fashion such that collisions between users are avoided. Noncontention channel access is usually implemented using some type of polling technique, wherein the central station sends a control message or synchronization signal to the remote stations as an indication for the remote to respond by transmitting data on the channel.

Using the well-known "explicit polling" technique, the central controller sends a polling signal to each remote station,

individually, to inquire if the remote has any information to send. A "poll list" of remote station addresses is used by the central controller to determine when a remote station is to be polled. If the polled remote station doesn't have a data message to send over the channel, the central controller goes on to poll the next remote. If the remote station does have a message to send, the data message is immediately transmitted over the channel in response to the poll. As used herein, the term "polling" includes the second-half of the procedure, wherein the polled stations return a message. Explicit polling has traditionally been considered rather inefficient, since each remote station has to wait for its individualized poll, establish bit and character synchronization, and then transmit its data message in response to the poll. Hence, a significant portion of the overall channel capacity is consumed by the polling signals themselves.

Another noncontention-based multiple-access protocol is referred to as "implicit polling." Under the implicit polling technique, each timing cycle on the channel is divided into a number of time slots, and a specific time slot within each cycle is reserved for a particular remote station. Each remote station, which is synchronized in time with the central station, is implicitly granted access to the channel during its individual time slot. In other words, the channel access is controlled by reserving time slots for each remote station to transmit, rather than being controlled by explicit polling signals from the central station.

In multipoint data communications networks using the implicit polling protocol, a fixed transmission time slot is reserved for each remote station in the network. Each time slot must be of a sufficient length to contain an entire data message packet. Hence, the channel is efficiently utilized only if each remote station has a data message to send during each cycle. If, however, only a few of the remote stations have messages to send during each cycle, then the channel remains idle during the preassigned time slots allocated to these non-responsive remote stations. When only a fraction of the remote users have data messages to send, an enormous amount of channel capacity is wasted in the empty time slots of an implicit polling system.

One advance over the prior art was provided by U.S. Patent No. 5,297,144 ("the '144 patent") assigned to the same assignee as the present invention. The '144 patent avoids some of the disadvantages of explicit and implicit polling by periodically allowing remote data stations to register a need for a data transmission with the central station under an implicit polling format. Registration is allowed under the '144 patent whenever the central station transmitted a reservation sync ("RS") frame. Contention was avoided following the RS frame by assigning different delay periods to each remote terminal for transmission of an access request following the RS frame.

Under the '144 patent a relatively fixed time period was allocated for the RS frame and access requests ("the reservation request period"). Following the reservation request period a second, variable length, time period is allowed for

polling the remote station and for transfer of data from the requesting remotes ("the polled data transfer period").

While the '144 patent ostensibly reduces power consumption within the remote stations through use of "sleep modes" such power savings is not practical where a remote station is to receive data from the control station. Under the teachings of the '144 patent, a remote station remains active during the polled data transfer mode (does not enter the sleep mode) only long enough to be polled and transfer data. Since data transfer from the central station to the remote station occurs at the end of the polled data transfer mode, and since the polled data transfer mode is of variable length, the sleep mode of the '144 patent cannot be used where data is to be transferred from the central unit to remote stations.

The '144 patent also allows for the addition of new remote stations to the relatively fixes reservation request period through the use of a "membership acquisition period". The membership acquisition period is a multiframe structure within the superframes after the polled data transfer period ("PDTP") wherein the central station accepts new remote stations (inserts new slots within the reservation request period). The membership acquisition period is a fixed time period within the superframe wherein a new remote station (or group of new remote stations) may seek to gain access to the communication system.

While the '144 patent has provided a significant advance over the prior art, the '144 patent still fails to provide a convenient method of coping within rapid membership changes. The '144 patent also fails to address the issue of

power savings where downlink transmissions must occur between the central station and remote stations. Because of the importance of power savings in portable data devices linked to a central station, a need exists for a method and means of remote station power control under dynamic loading conditions involving the two-way exchange of data between remote stations and the central station.

Summary of the Invention

An apparatus is provided for exchanging data between a central station and a plurality of wireless remote stations on a time divided communication channel. The apparatus includes means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol and means for polling during a second time period remote stations of the plurality of remote stations providing access requests under non-contention based protocols during the first time period.

The apparatus also allows remote stations to exchange data directly. Such direct exchange is possible where the central station acts to coordinate such exchanges while deferring the enablement of other users which may interfere on the communication channel.

Another aspect of the invention provides a second time period where data may be transferred from the central station to individual remote stations. A structure for broadcasting common

information to all remote stations is also provided within the second time frame.

Brief Description of the Drawings

The features of the present invention which are believed to be novel are set forth with particularity in the appended claims. The invention itself, however, together with further objects and advantages thereof, may best be understood with reference to the following description when taken in conjunction with the accompanying drawings, in which:

Figure 1 is a general block diagram of the wireless data communications network according to the present invention;

Figure 2 is a pictorial representation of the channel frame format utilized in the multiple-access signalling protocol of the present invention;

Figure 3 is a timing cycle diagram illustrating the two-stage reservation-based polling protocol and data exchange system of the present invention;

Figure 4A-C provides a summary of network control function by frame type in accordance with the invention along with a description of frame content within individual fields of the frame;

Figure 5 depicts a slot arrangement used within the request period in accordance with the invention;

Figure 6 is a timing cycle diagram similar to that of Figure 3 illustrating slot usage.

Figure 7 is a timing cycle diagram similar to that of Figure 6, wherein acknowledgement signals are returned to the remote stations after each data message;

Figures 8a and 8b are timing diagrams representing the power consumption of the remote station receiver and transmitter, respectively, during the reservation-based polling protocol timing cycle and data exchange of Figure 7;

Figure 9 is a detailed block diagram of one of the remote stations of the data communications network shown in Figure 1; and

Figure 10 is a detailed block diagram of the central station of the data communications network of Figure 1.

Detailed Description of the Preferred Embodiments

The solution to the problem of power savings in a dynamically loaded system requiring the two-way exchange of data between remote stations and a central station lies, conceptually, in mixing contention and non-contention based access protocols and in mapping a data transfer period into uplink and downlink epochs. The prior art has taught that either contention based protocols or non-contention based protocols may be used within access periods gaining entry to a multiple access system. Under the invention, it has been determined that an unexpected increase in efficiency may be achieved by using non-contention access protocols for remote stations requiring frequent data exchanges and contention access protocols to remote stations with less frequent data exchanges.

Mapping of data transfer periods, on the other hand, improves efficiency (reduces power consumption) by allowing an indicia of epoch locations to be transferred to remote stations at predetermined intervals. The indicia of epoch location may then be used by the remote stations to deactivate unnecessary power consuming devices during periods of inactivity.

Referring now to Figure 1, a general block diagram of a wireless multipoint data communications system 10 is shown. The system comprises a central station 12 and a number of remote stations 14. The central station 12 may be a stand-alone data processing and control entity or may be an access point (AP) used in conjunction with other data processors and systems over a larger hard-wired network.

Central station 12 communicates with remote stations 14 through an optical infrared transceiver 16 coupled to the central station via a hard-wired link 18. Each of the remote stations 14 includes an optical infrared transceiver which communicates with the central station by sending and receiving data messages over an infrared link 19. Depending upon the type of network, the central station may utilize the data messages itself, or route the data messages on to a different station in a local area network.

In the preferred embodiment, each of the remote stations is a portable, hand-held, battery-powered computer having an integrated infrared transceiver, as will be described in detail below. The remote stations also include a keypad for data input, and a display for data output. Although the present invention is particularly adapted for two-way communications over

a single-frequency infrared channel transmitting bursts of data packets in the half-duplex mode of operation, the present invention can also be used in full-duplex operation as well as half-duplex operation over single-frequency or split-frequency channels. In the preferred embodiment, infrared link 19 has a 4 Megabit data rate using Return To Zero with Bit Insertion (RZBI) encoding scheme. However, the present invention is not limited for use with only wireless links or the particular type of channel or data communications scheme shown here.

Figure 2 illustrates the specific channel frame format 20 used under the protocol for all information transfer and supervisory commands. The frame format of the invention basically follows the High-level Data Link Control (HDLC) data communications line protocol specification of the CCITT, or the Synchronous Data Link Control (SDLC) protocol specified by IBM. Hence, the published detailed specifications for the HDLC or SDLC protocols may be referred to for a further understanding of the common subject matter.

As shown in Figure 2, each frame is subdivided into a number of individual fields, wherein each field is comprised of a number of 8-bit bytes. The following paragraphs describe channel frame format 20:

<u>Preamble (PRE) 22</u>: This field is a 3-byte field whose purpose is to provide a means of establishing bit synchronization of the receiver with the received signal including the clock recovery signal. The value of the preamble is typically chosen to have a high content of transitions (e.g., "FFFFFF" because in RZBI encoding each "1" bit provides a high-low transition).

Start Delimiter (SD) 24: The purpose of the SD frame is to provide byte synchronization within the receiver. The 8 contiguous bits of the pattern provide a clear indication of the boundary between the "1" bits of the PRE and the bits of the SD. It is a unique "illegal" data structure because the bit insertion of the modulation scheme prevents this number of contiguous zero bits from occurring within the data (anyplace between the SD and ED fields).

Destination Identifier (DID) 26: This field contains the 2-byte address of the station to which the frame is being sent. In other words, in a polling frame, the DID field of a frame transmitted to a remote station first identifies the particular remote station being polled by the central station and then the DID field of a return frame identifies the central station as the destination for the data message being returned by the remote station. Each of the stations is assigned a unique identification code, or address. The remote stations typically receive a new DID address each time the remote station registers with the network 10. However, a dynamic address determination procedure could also be used. In the preferred embodiment, the addresses of remote stations (non-controller stations) begin with hex and increase to the maximum amount of remote stations allowed in the network (e.g., 7FFF hexadecimal). Controller stations (e.g., central station 12) may be assigned other numerical values (e.g., 8000-EEED hexadecimal). A value of FFFF hex in this field denotes a broadcast frame, which would be received by all stations.

Source Identifier (SID) 28: This field is the 2-byte address of the station sending the frame. To ensure the integrity of the data being transmitted, both the destination and source addresses are included within each frame.

Type of Field (TYP) 30: The 1-byte field indicates to the receiver how to interpret the frames contents and in effect provides a control function. A summary of the possible types of frames are as follows: RSYNC, MRSYNC, RegRTS, RTS, FORF, DSYNC, EDSYNC, RegCTS, CTS, DATA, MDATA, and ACK. The meaning and content of the types of frames listed may be best understood by reference to FIGS. 4A-C. The use of the frames may be best understood by reference to subsequent sections.

Control Flags: This is a 1-byte control field containing bit-mapped flags, primarily used for supervising commands. In the preferred embodiment, control field 32 includes priority flags and retransmissions flags, which will be described below.

Information (INFO) 34: This is a variable length field used for transferring data. The INFO field 34 is also used in conjunction with certain types of frames (e.g., RSYNC, MRSYNC, DYSNC, and EDSYNC) as a repository for an indicia of epoch location (e.g., the location of upward data transfer period (upward period), broadcast period and downward data transfer period (downward period) within the overall data exchange period (data period)).

Frame Check Sequence (FCS) 36: This 4-byte field is used to detect bit errors which may occur during transmission. In the present embodiment, a 32-bit cyclic redundancy check (CRC)

algorithm is used to detect errors in fields 26, 28, 30, 32, and 34.

<u>End Delimiter (ED) 38</u> and <u>Postamble (Post) 40</u>: The purpose of the ED 38 is to allow the receiver to detect an end of frame. The purpose of the POST 40 is to prevent the receiver from mistaking an ED/POST combination for an SD/DID combination in that the hexadecimal value of 0EEEE could be an invalid DID.

Figure 3 illustrates a repeating frame structure (superframe) used by the system 10 to exchange information between the central station 12 and the remote station 12. Each frame making up the superframe has the frame format described above.

Superframes are not always of the same temporal length. The superframe, in turn, may be divided into a variable length period used for receipt of access requests (request period) 50 and a variable length field used for data exchanges (data period) 51.

The central station 12 identifies the beginning of the superframe to the remote stations 14 by transmission of a request synchronization (RSYNC) frame or a mandatory request synchronization (MRSYNC) frame 52. (The RSYNC frame requires only those remote stations 14 desiring access to respond while the MRSYNC requires all remote stations 14 to respond.) The remote stations 14 identify the RSYNC or MRSYNC frames by reference to the type field of the frame (FIG. 4A-C). In addition to identifying the beginning of the superframe, the RSYNC or MRSYNC frame 52 provides information within the INFO field 34 (FIG. 4A) relative to the number and type of slots (slots using a non-contention

based access protocol or a contention based access protocol) within the request period 50. The slot information is used by the remote stations to facilitate system access (to be explained later) or to power-down during the request period 50 if the remote station 14 does not need access to the network 10.

Following the request period 50, the network 10 enters a data period 51. The central station 12 identifies the beginning of the data period 51 to the remote station 14 by transmission of a data descriptor frame 53 (e.g., a data synchronization (DSYNC) or extended data synchronization (EDSYNC) frame). Contained within the INFO field 34 (FIG. 4A) of the DSYNC or EDSYNC frame 53 is temporal information relative to the length of each subsection of the data period 51. The temporal information, as above, is used by the remote stations 14 to reduce a duty cycle of activation by powering-down during appropriate portions of the data period 51.

In accordance with an embodiment of the invention, the slots of the request period are divided into two groups where a first group of slots allows for random access under a contention based protocol (contention slots) and a second group of slots allows for access under a non-contention protocol (reserved slots) (e.g., under an implied polling protocol). Under the invention, the number of contention slots may be constant or may vary based upon an estimate of the number of unregistered remote stations within the service coverage area of the network 10. The number of reserved slots, on the other hand, is adjusted based upon loading. When a remote station 14 is first activated the remote station 14 is granted access to the network 10 under a

two-step process. First the remote station 14 transmits an access request within a contention slot. The central station 12 upon receipt of the access request within the contention slot then, as a second step, assigns the remote station 14 to a non-contention slot before finally granting access.

The remote station 14 first monitors for a RSYNC or MRSYNC frame 52. Since the remote station 14 does not yet have a reserved slot, the remote station 14 must access the network 10 through a contention slot. The remote station 14 identifies contention slots by examining the contents of the INFO field 34 of the RSYNC or MRSYNC frame 52. Contained inter alia within the INFO field 34 of the RSYNC or MRSYNC frame (FIG. 4A) is the total number of slots in the request period and the total number of reserved slots. By knowing the location of the reserved and contention slots relative to the RSYNC or MRSYNC frame (e.g., the non-contention slots may immediately follow the RSYNC or MRSYNC frame), the remote station 14 can determine the location of the contention slots. Access may then be secured through a randomly selected contention slot.

By way of example, FIG 5 depicts a request period having 10 slots. If the reserved slots were designated as being slots 1-7, then slots 8-10 would be the contention slots. An INFO field 34 of a RSYNC or MRSYNC frame 52 in such a case would indicate a total slot number of 10 and a total reserved slot number of 8. Using known methods, the remote station would then randomly generate a number in the range of 1 to 3 and add the randomly selected number to 8 for a final determination of the contention slot to be used in requesting access.

In requesting access to the network 10, the remote station 14 sends a registration request to send (RegRTS) frame (FIG. 4B) within the selected contention slot. The INFO field 34 of the RegRTS frame contains a 48 bit address of the requesting remote station 14 along with coding within the type field that the frame is a RegRTS frame.

Upon receipt of the RegRTS from the remote station 14 by the central station 12, the central station 12 verifies by reference to a memory (not shown) that the address of the remote station 14 that the station is authorized to access the network 10 and that the remote station 14 has a software version compatible with the network 10. Upon verifying that the remote station 14 is an authorized user and is compatible with the network 10, the central station 12 issues a local identifier in favor of the remote station 14. The central station 12, on the other hand, does not immediately transmit the local identification to the remote station. Under the invention the central station waits until the next downward portion of the data period 51 before transmitting the identifier to the requesting remote station 14.

Contained within the local identifier is an identifier of a reserved slot of the request period 50 allocated for use by the remote station 14. The central station 12 may create a reserved slot for the remote station 14 by expanding the length of the request period to 11 slots or may assign the remote station 14 to an unoccupied slot of reserved slots 1-8 (FIG. 5).

Likewise, the central station 12 may de-allocate a slot previously reserved for use by other remote stations 14 based on

certain operating parameters. The central station 12 may de-allocate slots for instance where the time since the last use of the slot exceeds some threshold value or if the remote station 14 does not respond to a known number of consecutive MRSYNC frames.

During the next downward period of the data period 51 the central station 12 transmits the local identifier to the remote station 14 through use of a registration clear to send (RegCTS) frame (FIG. 4B). Upon receiving the RegCTS, the remote station retrieves the local identifier and, using the retrieved local identifier, may transmit a Request to Send (RTS) within the designated reserved slot under an implicit polling format during the request period 50 of the next superframe.

Under an alternate embodiment, the remote station 14, upon receipt of a RegCTS may immediately respond by transmitting data. Alternately, a central station 12 may transmit a RegCTS at any time to fill "holes" in the request period (e.g., when a remote station 14 is deactivated or leaves the service coverage area of the network 10).

In general, implicit polling is performed during the request period 50, and explicit polling -- of only those remote stations which requested access to the channel -- is performed during the data period 51.

To initiate the superframe, the central station broadcasts an RSYNC or MRSYNC frame 52 to all the remote stations. The RSYNC or MRSYNC frame is issued periodically, and it defines the start of a number of time slots of the request period. In the preferred embodiment, the central station sends

a RSYNC or MRSYNC frame at least once every second. If there is less data to exchange then the superframe would occur more often, but not more often than once every 100 ms. If there were less data than could be transferred within the 100 ms interval, then the communication channel would be idle for a portion of the 100 ms.

Under an alternate embodiment, an RTS of the remote station 14 specifies the number of data frames it wants to send during the superframe. It is then up to the central station 12 to determine how many times a remote station 14 gets polled. For instance, a central station 12 wouldn't let an entire superframe be "eaten up" by a single station if it requests to be polled too often. Once a request period 50 is complete, the central station 12 has a picture of all upward and downward data periods, and it will divide up the superframe equitably.

A central station 12 may indicate within the RTS frame during the RTS/CTS/DATA/ACK sequence how many frames it will send to the remote station 14 during a superframe. During a DATA/ACK sequence, the use of a "more" bit indicates to the remote station 14 that there is more data to be transmitted during the superframe.

Every remote station has a preassigned waiting period that will begin upon the reception of the RSYNC or MRSYNC frame. These waiting periods are illustrated as time slots TS in Figure 6, which fill up the remainder of the request period 50.

Since remote station 1 has been assigned the first time slot, it issues a reserved slot request RR frame 54 if it has data to transmit on the channel. Hence, the first time slot has

been replaced with the reserved slot request frame RR_1 (RTS frame) transmitted from remote station 1. As seen in the example of Figure 6, no reserved slot request frame was issued in time slot 2 (frame 55), and a reserved slot request frame RR_3 was issued from remote station 3 in time slot 3 (frame 56). In the example shown, a maximum number $X-X_c$ (where X is total slots and X_c is contention slots) denotes the number of active remote stations in the network, and, accordingly, the number of preassigned time slots. (See frame 56.) Note that, in this example, the absence of a reserved slot request frame in a time slot represents a negative access request signal to the central station 12. As will be seen below, an alternate embodiment of the protocol always returns either a positive or negative access request signal to the central station upon issuance of a MRSYNC frame.

After every station has been given a chance to make a reservation, the central station will switch to a modified explicit polling mode, wherein it will sequentially issue a CTS frame to every remote station 14 that made a reservation.

Before the central station 12 begins the explicit polling, on the other hand, the central station 12 must describe the data period 50 for the benefit of those remote stations 14 that may wish to power-down for portions of the data period 50. The central station 12 describes the data period 50 to the remote stations 14 by transmitting a DSYNC or EDSYNC frame 53. (The DSYNC and EDSYNC frames differ primarily in the amount of information provided. In general, the EDSYNC allows for a lower duty cycle of remote stations 14).

If either a DSYNC or an EDSYNC frame 53 is used, then the reader will find via reference to FIG. 4A that the length of the polling period for the upward transmission of data is to be found within the INFO field 34 of the DSYNC or EDSYNC frame 53. A remote station not needing to transfer data to the central station 12 may use the time period specified to deactivate its transmitter and receiver until a point just before the broadcast period, where the remote station 14 must again re-activate its receiver for the receipt of system information during the broadcast period.

As illustrated in Figure 6, the central station polls the first remote station during frame 60 of the upward period with CTS frame P_1, since remote station 1 sent its reserved slot request frame RR_1 during frame 54. Immediately upon receiving the poll signal addressed to remote station 1, that station responds with its data packet $DATA_1$ during frame 62. The central station then checks its poll list to determine which remote station is to be polled next. In the example shown, remote station 3 is polled via poll frame P_3 during frame 64, and it responds with its data packet $DATA_3$ during frame 66. The polling ends upon the completion of the response of the last station on the list, which, in this case, was remote station 3.

Priority message capability is also provided for in the reservation-based polling and data exchange protocol of the present invention. Recall that the control field 32 of the channel frame format 20 (FIG. 2) includes a number of bit-mapped priority flags. In the preferred embodiment, four levels of priority can be implemented using two priority flag bits. If any

remote station had a priority message to send, then that station would set its priority flags to the appropriate priority level, and transmit a reserved slot request RR frame to the central station in its preassigned time slot during the reserved slot request period. Upon receipt of this reserved slot request frame containing priority information, the central station would rearrange the poll list into priority-level order. Accordingly, the central station would poll the remote stations in priority-level order.

The timing cycle diagram shown in Figure 6 can be used to illustrate the reservation-based polling protocol with priority-level polling. Assume that the time slots TS_1, TS_2, TS_3, (frames 54-56) of the reserved slot request period are sequentially assigned to correspond with three remote stations 1-3. If all three remote stations had non-priority messages to send, then each would send its reserved slot request RR frame during the appropriate time slot, and the central station would poll each remote station in numerical order, i.e., the poll list would appear as: P_1, P_2, P_3. If, however, remote station 3 had a level-one priority message to send, and remote station 2 had a level-two priority message to send, then these stations would indicate such using the priority flags in the control fields of their reserved slot request frames. The central station would then re-order its poll list to appear as: P_3, P_1, P_2. Thus, the remote stations are polled in priority-level order. Numerous other multiple-level priority message schemes can be used with the present invention, a few of which will be described below.

Figure 7 represents a similar timing cycle diagram to that of Figure 6, with the addition that an acknowledgement (ACK) frame is transmitted from the central station to the remote station after the reception of each data message from the remote station. In order to send an ACK frame, the central station 12 must, first, correctly receive the data message before issuing an ACK frame (FIG. 4C).

The example of Figure 7 illustrates that, during the reservation request period, remote stations 1 and 3 have transmitted reserved slot request frames 54 and 56, respectively. Therefore, during the upward data transfer period, each of these two remote stations is polled. As before, a first poll frame P_1 is issued from the central station in frame 60, and data packet $DATA_1$ from remote station 1 is returned during frame 62. However, now an acknowledgement frame AK_1 is sent from the central station to remote station 1 during frame 63. A similar polling/data transfer/acknowledgement sequence occurs for remote station 3 during frames 64, 66, and 67. As only partially shown in Figure 7, remote station $X-X_c$ was polled, it transmitted its data packet, and its acknowledgment frame AK_X is shown being returned during frame 69.

If the remote station 14 does not receive an acknowledgement (ACK) from the central station 12 following a data transfer (or does not get polled), then the remote station 14 sends a reserved slot request (RR) during the next request period 50. If the remote station 14 does not get a response after 3 tries, the data is discarded.

The broadcast period follows the upward period. Any stations which may have de-activated during the upward period must re-activate for the broadcast period. During the broadcast period, data is broadcast from the central station 12 to all remote stations 14. Data frames (FIG. 4C) during the broadcast period are sent with the broadcast DID (e.g., FFFF hexadecimal). Broadcast data frames are not preceded by an RTS/CTS exchange and are not acknowledged by receiving remote stations 14. If there is no broadcast data to be sent from the central station 12 to the remote stations 14, then an EDSYNC frame 53 at the beginning of the data period 51 may be used to indicate a broadcast length of zero.

Following the broadcast period is the downward data period. If the data descriptor 53 at the beginning of the data period 51 were a DSYNC frame, then all remote stations 14 must remain activated during the downward data period.

If, on the other hand, the data descriptor 53 were a EDSYNC frame, then the contents of the EDSYNC would provide advance notification of which remote station(s) 14 would receive data and, therefore, which remote stations 14 would remain activated during the downward data period. Other remote stations 14 not present within the list of the EDSYNC frame may deactivate for the duration of the downward data period.

Data transfer from the central station 12 to the remote stations 14 during the downward period may occur under either of two possible scenarios. The central station may either transmit an RTS and wait for a CTS before transmitting the data, or may simply transmit a data frame and wait for an acknowledgement

response. The use of the RTS by the central station 12 avoids the unnecessary transmission of data when the remote station 14 may not be within range of the central station 12. The use of the RTS/CTS exchange, on the other hand, causes more overall data traffic between the central station 12 and remote station 14.

If the remote station received an erroneous data message, then a negative acknowledgment frame would be returned to the central station. If the central station received neither an acknowledgement frame nor a negative acknowledgement frame from the remote station, then the central station would retransmit the same data message in the next superframe.

Where the RTS/CTS/DATA/ACK sequence is used and there is no response to the RTS, or if the CTS is received with errors, or if after the RTS/CTS/DATA sequence, the ACK isn't received, or if the ACK is received with errors, then the central station 12 begins its retransmission with the retry bit of the RTS frame set. On the other hand, where the DATA/ACK sequence is used and there is no ACK received, or if the ACK is received with errors, then the central station begins its retransmission with the retry bit of the DATA frame set.

Depending upon the requirements of the particular data communication system, it may be advantageous for the central station to track and report on the number of active remote stations in the network -- whether or not each remote station has a data message to send. For this purpose, the central controller would issue a mandatory request synchronization (MRSYNC) frame to all of the remote stations. When a remote station receives this frame, it responds with a RTS frame if it has data to send,

or it responds with a forfeit (FORF) frame if it does not. If a particular remote station does not respond to the global reservation sync frame, then the central station assumes that the particular remote station 14 is not presently active. In this manner, all of the active remote stations will be accounted for by the system without affecting the throughput of the data communications channel.

Periodically, the central station issues a frame (RSYNC, MRSYNC, DSYNC, or EDSYNC) including a superframe number. The superframe number may be used by the remote stations 14 as a functional check of proper operation (e.g., that a particular sleep mode interval did not cause a remote station 14 to miss part of a superframe).

The timing diagrams of Figures 8a and 8b illustrate the sleep mode of remote station 3. During the sleep mode, the controller in the remote station 14 may disable the infrared transmitter and/or receiver circuitry, as well as any other circuitry such as a communications processor which is not being used at the time. This sleep mode ensures minimum power consumption to extend the life of the battery. Figure 8a represents the power consumption of the remote station receiver, and Figure 8b represents the power consumption of the remote station transmitter. These two timing diagrams correspond to the timing cycle shown in Figure 7, wherein acknowledgment frames are utilized.

Since the reservation sync frames 52 and descriptor frames 53 are substantially periodic, the remote station can be programmed to periodically enable its receiver to wait for a

reservation sync frame 52 and descriptor frame 53. Accordingly, as shown in Figure 8a, the receiver of remote station 3 is turned on at time t_0, which precedes the occurrence of the reservation sync frame RS at time t_1 by a sufficient amount to account for clock tolerances. After the reservation sync frame has been received, the receiver is disabled at time t_2.

At time t_3, the transmitter circuitry is enabled such that the reservation request frame RR_3 can be transmitted during time slot 3. At time t_4, the transmitter returns to the sleep mode. At time t_5, the reservation request period has ended, and the polled data transfer period (upward period) has begun.

In general, if the remote station 14 has requested access to the network than at least the receiver needs to remain active during the upward period for the receipt of polling messages. Upon receipt of a polling message, directed to the remote station, the receiver may be deactivated and the transmitter activated. Also, if the descriptor for the data period 51 is a DSYNC frame, then the remote station 14 must remain active for the broadcast period and for the downward period. Further, if a frame directed to the remote station 14 is detected by the remote station 14, then the transmitter of the remote station 14 must be activated for transmission of acknowledgement message.

If the descriptor 53 of the data period 51 is a EDSYNC frame, then the remote station 14 shuts down unless otherwise required. If the remote station 14 has transmitted an access request during the request period, then the receiver of the remote station 14 would remain active until polled, at which time the receiver would deactivate and the transmitter activate for

transmission of the data frame. At the end of the data frame the transmitter would again deactivate and the receiver activate for receipt of the acknowledgement frame from the central controller 12. Likewise, the remote station would only activate for the broadcast period if the EDSYNC message indicated that the broadcast period would have a non-zero time period, or if a data frame were to be directed to the mobile station 14 during the downward period.

Accordingly, remote station 3 (FIG. 8a) must enable its receiver such that it can wait for its poll frame P_3. At time t_6, the poll P_3 has been received such that the receiver can be disabled. However, the transmitter is immediately enabled since data packet $DATA_3$ must be transmitted during frame 66. From times t_7 to t_8, acknowledgement frame AK_3 is being received by remote station 3. After time t_8, the receiver of the remote station can return to its sleep mode until the broadcast period and downward period. Where a DSYNC descriptor 53 is received and if no messages were received by the remote station 3 (as depicted in FIG. 8a) (under either DSYNC or EDSYNC descriptors 53), then at least the transmitter will remain deactivated until the next superframe. As can now be seen, the sleep mode is used by the remote station to conserve battery power when the central station 12 is communicating with other remote stations 14. Various other sleep mode configurations may also be used, particularly since many of the communications processors used in the remote stations may include their own internal power conservation circuits and software.

Referring now to Figure 9, a detailed block diagram of one of the remote stations is shown. As described above, each remote station 14 includes a transceiver which communicates with the central station via an optical-infrared data link 19. The heart of the remote station is a remote controller 110 which, in the preferred embodiment, is a Motorola 68EC000, available from Motorola Corporation, operating at 8 Mhz. Remote controller 110 interfaces with a data processor 112 and a communications processor 114, such that data processor 112 can communicate over the infrared link using the polling protocol described above. In the preferred embodiment, data processor 112 may be part of an EPSON Model No. H1001BEW hand-held computer, and communications processor 114 may be an 82590 LAN interface chip also available from Intel or may be a Field Programmable Gate Array (FPGA) with custom programmed logic provided by Spectrix Corp., of Evanston Illinois.

Communications processor 114, in turn, controls an infrared transmitter 116 and an infrared receiver 118. Infrared transmitters and receivers are well-known in the art. In order to perform the control of the sleep mode for the remote station, remote controller 110 also controls the application of power from power supply 124 to the transmitter and receiver blocks. In the preferred embodiment, power supply 124 is contained within the hand-held computer of the remote station 14. A clock 126 and a memory 128 are also connected to remote controller 110 in order to perform the synchronization and station identification functions of each remote station.

Figure 10 is a detailed block diagram of central station 12 of the data communications network shown in Figure 1. In order to communicate with the remote stations, the central station includes an external transceiver 16. In the preferred embodiment, infrared transceiver 16 is located at a distance from central station 12, since a personal computer is used for the network controller and since the infrared link must be direct line-of-sight. A network controller 130 interfaces an input/output port 132 to a communications processor 134 such that the reservation-based polling protocol of the present invention is used to transmit and receive data from infrared link 19 to I/O port 132 via infrared transmitter 136, infrared receiver 138, and hard-wired link 18. In the preferred embodiment, the function of network controller 130 is performed by an IBM-compatible personal computer using a DOS-based operating system. The personal computer typically includes a memory 140, a clock 142, a display 144, and a keyboard 146.

In review, it can now be seen that the present invention provides an improved contention and noncontention-based multiple-access signalling protocol for a data communications network which efficiently utilizes a single channel even when only a fraction of the users have data messages to send at a given time. The reservation-based polling protocol is particularly adapted for use with a large number of portable battery-powered computer devices communicating with a central station via an infrared link.

While specific embodiments of the present invention have been shown and described herein, further modifications and

improvements may be made by those skilled in the art. All such modifications which retain the basic underlying principles disclosed and claimed herein are within the scope of the invention.

What is claimed is:

1. An apparatus for exchanging data between a central station and a plurality of wireless remote stations on a time divided multiple access communication channel comprising: means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol; and means for polling for data transfers during a second time period the remote stations of the plurality of remote stations providing access requests under non-contention based protocols during the first time period.

2. The apparatus as in claim 1 wherein the means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol further comprises a plurality of time division multiple access slots within the first time interval.

3. The apparatus as in claim 1 wherein the means for receiving access requests from remote stations of the plurality of remote stations during a first time interval under a contention based protocol and a non-contention based protocol further comprising means, located within a slot of the plurality of slots, for identifying contention slots and non-contention slots to the plurality of remote stations.

4. The apparatus as in claim 3 further comprising means for receiving an access request from a remote station of the plurality of remote stations within an identified contention

slot and allocating a non-contention slot to the requesting remote station.

5. An apparatus for reducing a duty-cycle of activation of a remote station exchanging data with a central station on a wireless time divided multiple access communication channel comprising: means for receiving an access request from the remote station during a first time interval under one of a contention based protocol and a non-contention based protocol; means for exchanging data during a second time period with the remote station providing the access request under the non-contention based protocol during the first time period for data transfers; and means for providing an indicia of epoch length of the first and second time periods to the remote station by the central station.

FIG. 1

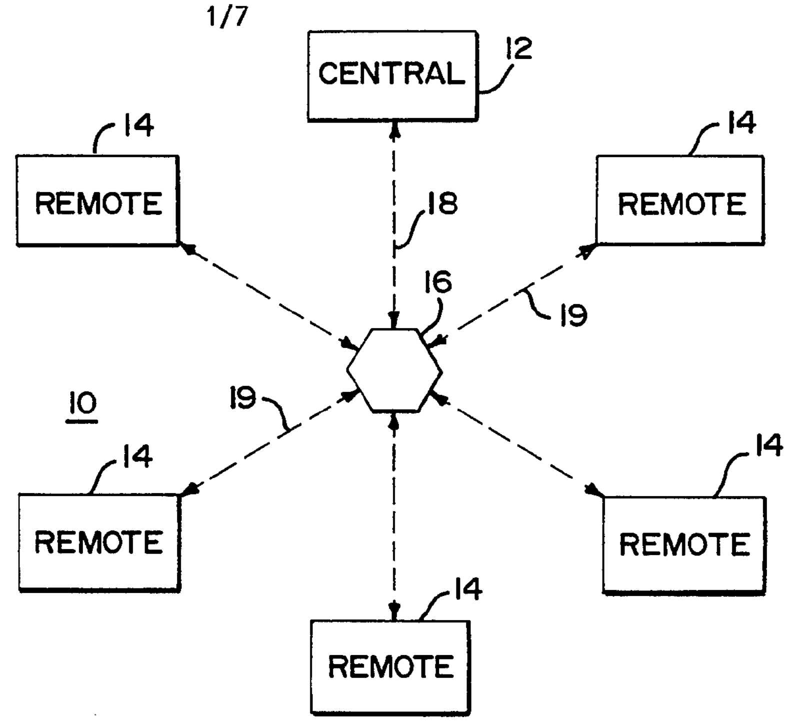


FIG. 2

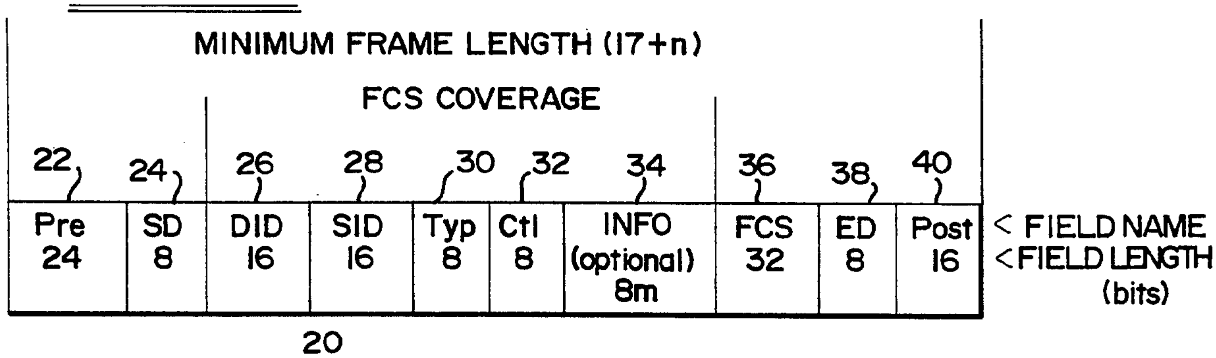


FIG. 3

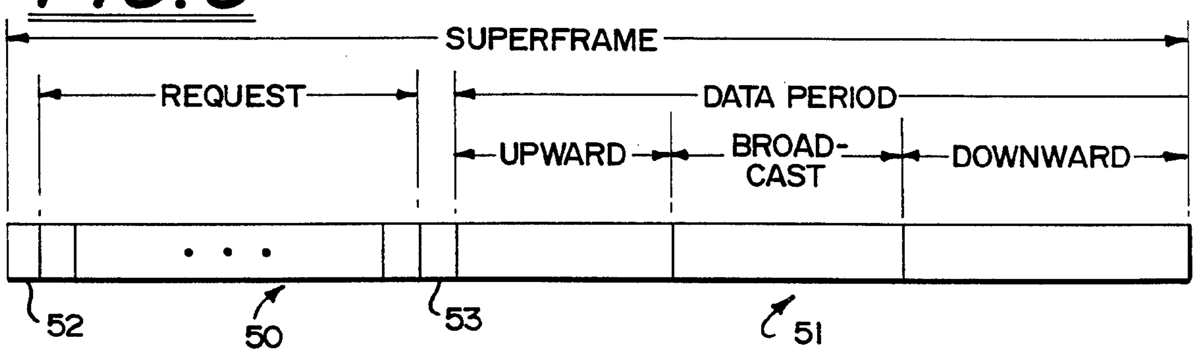


FIG. 5

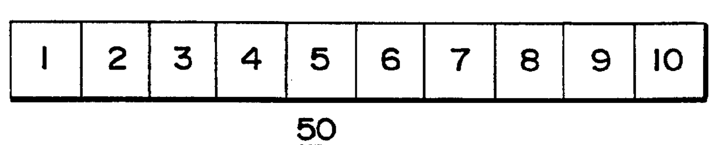


FIG. 4a

RSYNC-Request Sync
MRSYNC-Mandatory Request Sync

	Preamble	
	SD	
	DID	< Broadcast
	SID	< ID of originating controller
	Type	< RSYNC or MRSYNC
	Control	< Control flags: AP
	Version	< Software version
INFO	TotalSlots	< Total number of time slots in SyncPeriod (including RegSlots)
	RegSlots	< Total number of time slots which are for registration only
	SuperFrame	< Superframe number
	FCS	
	ED	

DSYNC-Data Sync

	Preamble	
	SD	
	DID	< Broadcast
	SID	< ID of originating controller
	Type	< DSYNC
	Control	< Control flags: AP
INFO	UpLength	< length of Upward Data Period
	SuperFrame	< Superframe number
	FCS	
	ED	

EDSYNC-Extended Data Sync

	Preamble	
	SD	
	DID	< Broadcast
	SID	< ID of originating controller
	Type	< DSYNC
	Control	< Control flags: AP
	UpLength	< length of Upward Data Period
	BCastLength	< length of Broadcast Data Period
INFO	DownLength	< length of Downward Data Period
	ListLength	< length of following list
	List	< list of stations for which there is downward data in this superframe
	SuperFrame	< Superframe number
	FCS	
	ED	

FIG. 4b

RegRTS-Registration Request

	Preamble	
	SD	
	DID	< Destination Station ID
	SID	< registration slot number (temporary station ID)
	Type	< regRTS
	Control	< Control flags: none used
INFO	Version	< Software version
	SA	< Address station registering, 48-bit address
	FCS	
	ED	

RTS-Request To Send

	Preamble	
	SD	
	DID	< Destination Station ID
	SID	< Source Station ID
	Type	< RTS
	Control	< Control flags: AP, sequence, out-of-sequence, retry, more
INFO	DataCount	< Number of data frames station wants to send to DA
	DataLength	< Length, in octets, of data the source wants to send
	DA	< Address station to which data is to be sent, 48-bit address
	FCS	
	ED	

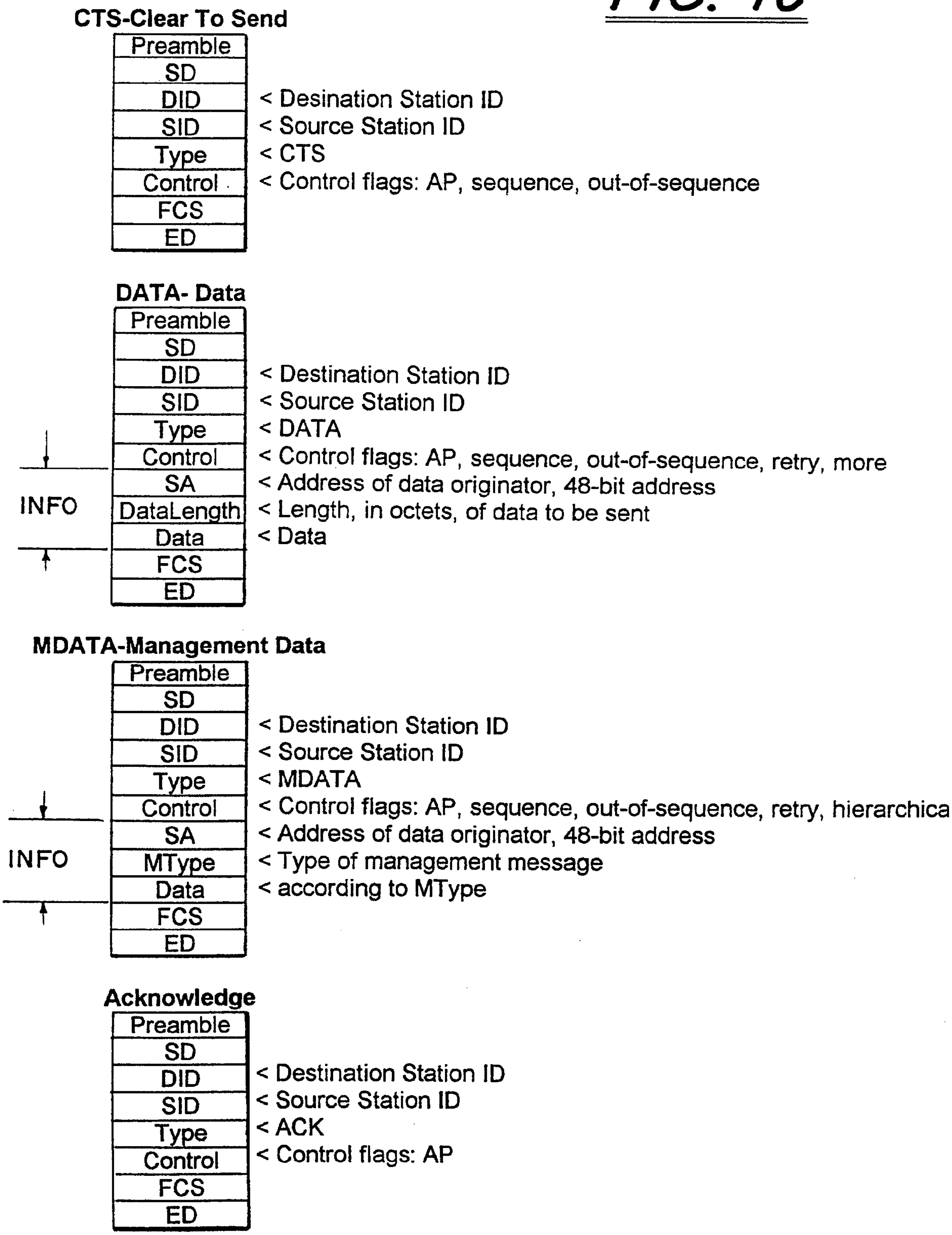
FORF-Forfeit

Preamble	
SD	
DID	< Destination Station ID
SID	< Source Station ID
Type	< FORF
Control	< Control flags: none used
FCS	
ED	

RegCTS-Registration Clear to Send

	Preamble	
	SD	
	DID	< Destination Staion ID
	SID	< ID of originating controller
	Type	< ReCTS
	Control	< Control flags: AP
INFO	ID	< ID assigned to station
	SA	< Address station registering, 48-bit address
	FCS	
	ED	

FIG. 4c


CTS-Clear To Send
Preamble
SD
DID
SID
Type
Control
FCS
ED
< Desination Station ID
< Source Station ID
< CTS
< Control flags: AP, sequence, out-of-sequence
DATA- Data
Preamble
SD
DID
SID
Type
Control
SA
DataLength
Data
FCS
ED
INFO
< Destination Station ID
< Source Station ID
< DATA
< Control flags: AP, sequence, out-of-sequence, retry, more
< Address of data originator, 48-bit address
< Length, in octets, of data to be sent
< Data
MDATA-Management Data
Preamble
SD
DID
SID
Type
Control
SA
MType
Data
FCS
ED
INFO
< Destination Station ID
< Source Station ID
< MDATA
< Control flags: AP, sequence, out-of-sequence, retry, hierarchical
< Address of data originator, 48-bit address
< Type of management message
< according to MType
Acknowledge
Preamble
SD
DID
SID
Type
Control
FCS
ED
< Destination Station ID
< Source Station ID
< ACK
< Control flags: AP

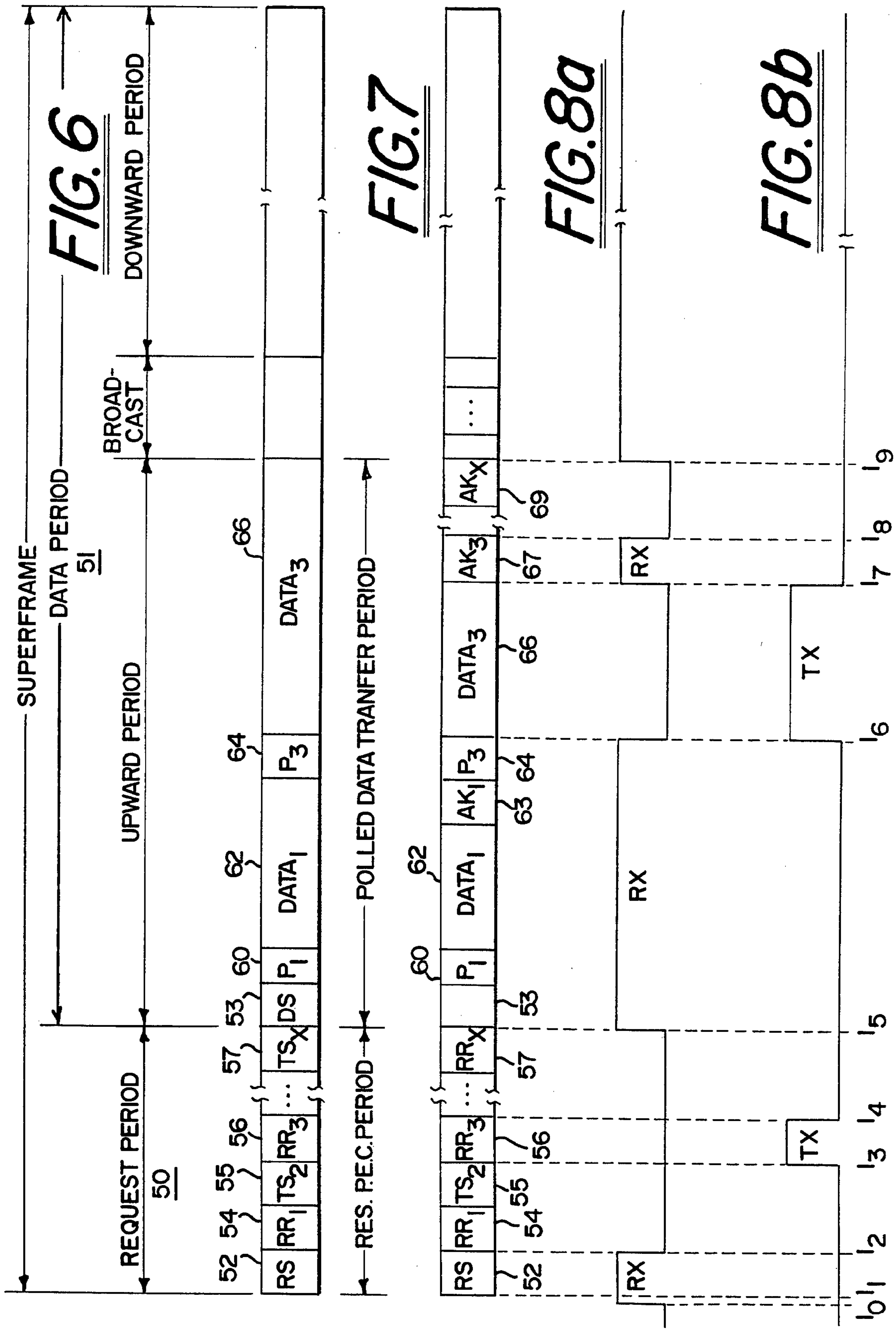


FIG.6
SUPERFRAME
DATA PERIOD
51
REQUEST PERIOD
50
UPWARD PERIOD
BROAD-CAST
DOWNWARD PERIOD
RS RR1 TS2 RR3 ... TSX DS P1 DATA1 P3 DATA3
52 54 55 56 57 53 60 62 64 66
FIG.7
RES. P.E.C. PERIOD
POLLED DATA TRANFER PERIOD
RS RR1 TS2 RR3 ... RRX P1 DATA1 AK1 P3 DATA3 AK3 AKX ...
52 54 55 56 57 53 60 62 63 64 66 67 69
FIG.8a
RX RX RX
FIG.8b
TX TX
t0 t1 t2 t3 t4 t5 t6 t7 t8 t9

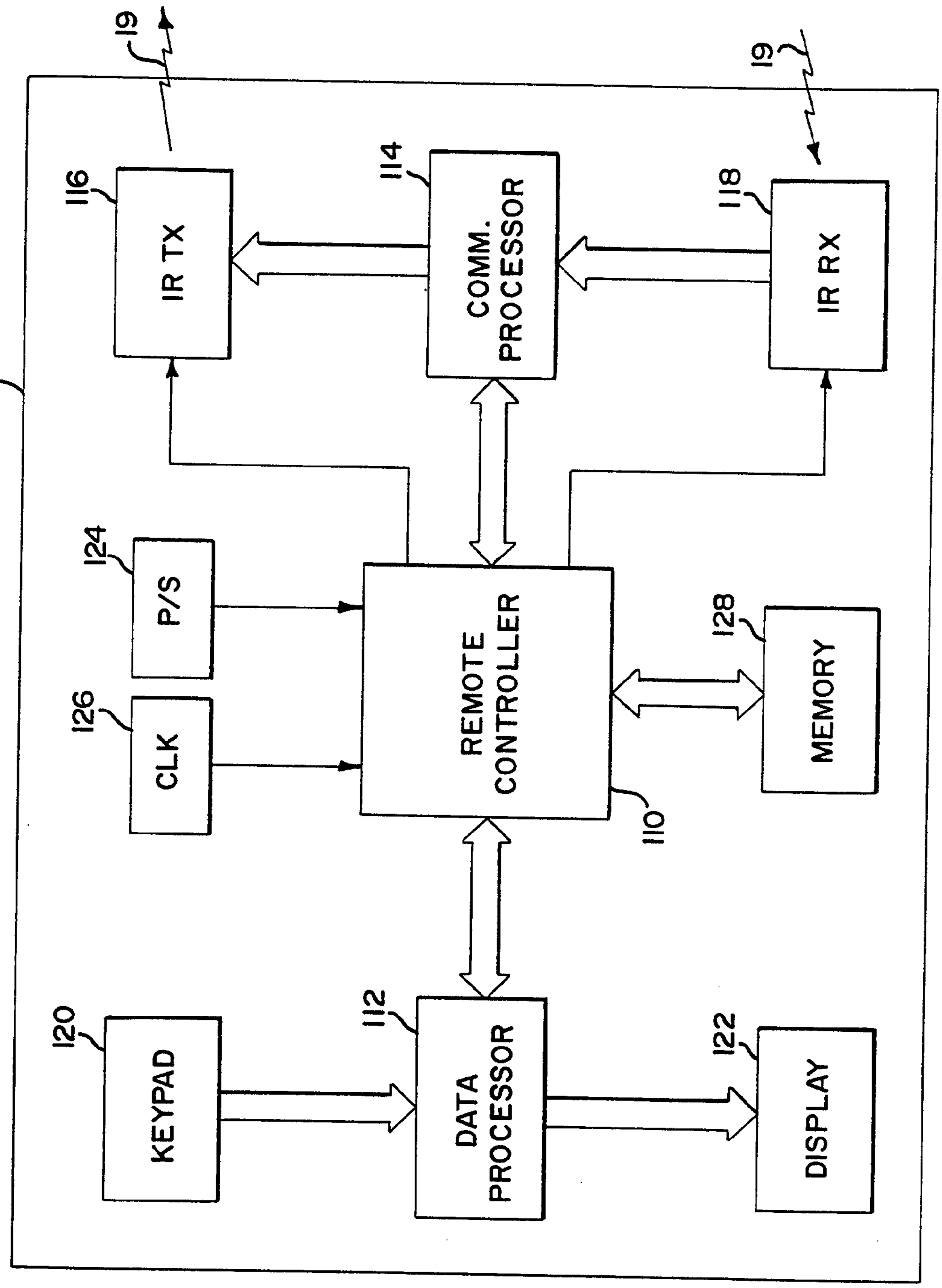

19
19
14
116
IR TX
114
COMM. PROCESSOR
118
IR RX
124
P/S
126
CLK
REMOTE CONTROLLER
110
128
MEMORY
120
KEYPAD
112
DATA PROCESSOR
122
DISPLAY

FIG. 9

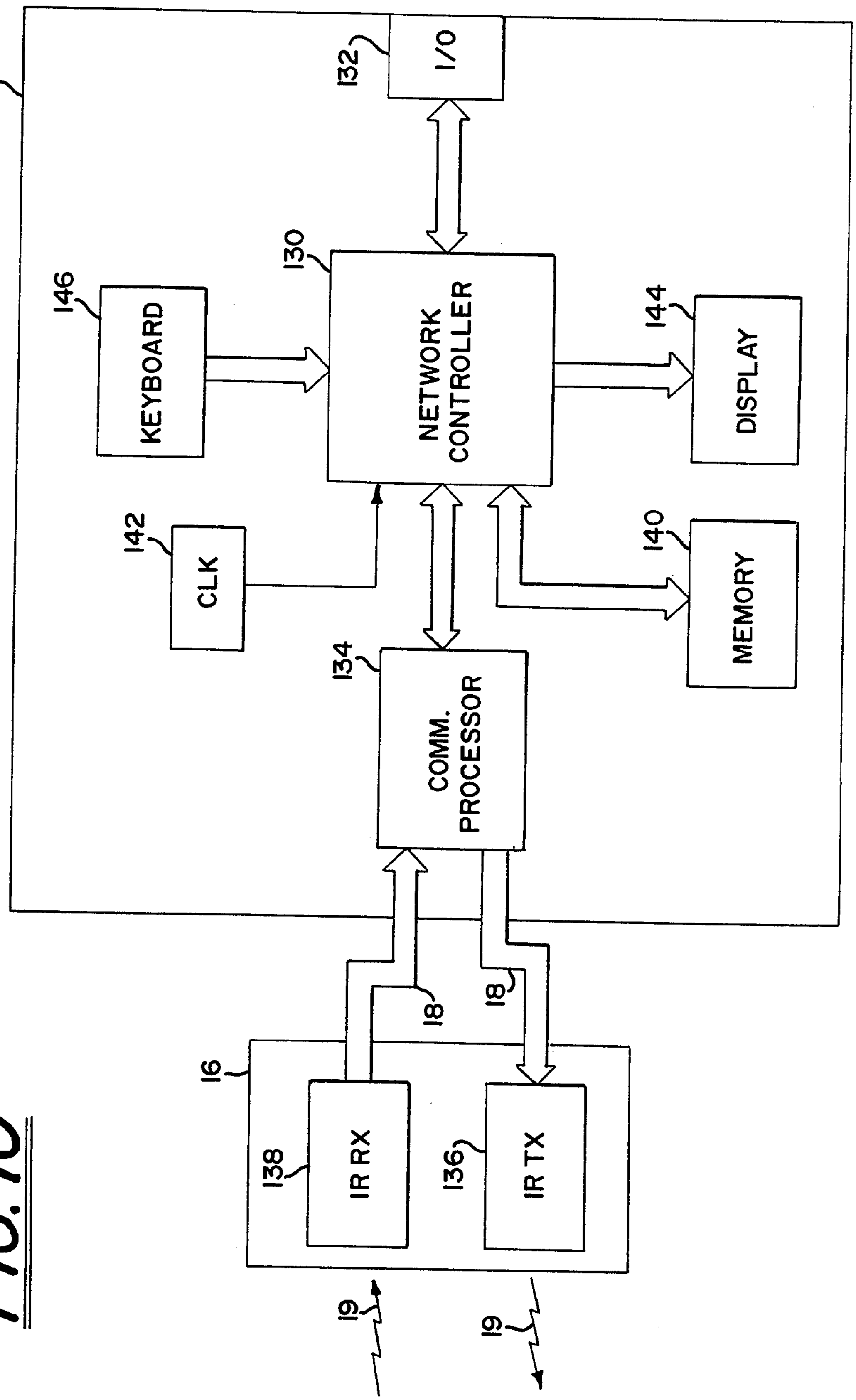

12
I/O
132
130
NETWORK CONTROLLER
146
KEYBOARD
144
DISPLAY
142
CLK
140
MEMORY
134
COMM. PROCESSOR
18
18
16
138
IR RX
136
IR TX
19
19

FIG. 10

INTERNATIONAL SEARCH REPORT

International application No.
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A. CLASSIFICATION OF SUBJECT MATTER

IPC(6) : H04J 3/16
US CL : 370/095.200 , 095.300 , 085.800 , 085.200 ; 340/825.080
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 370/095.200 , 095.300 , 085.800 , 085.200 , 085.700 , 095.100 ; 395/200.090 ; 455/004.100; 340/825.080

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

NONE

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

NONE

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US, A, 5,297,144 (GILBERT ET AL.) 22 March 1994, columns 1-6 and figures 1-4.	1-5

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

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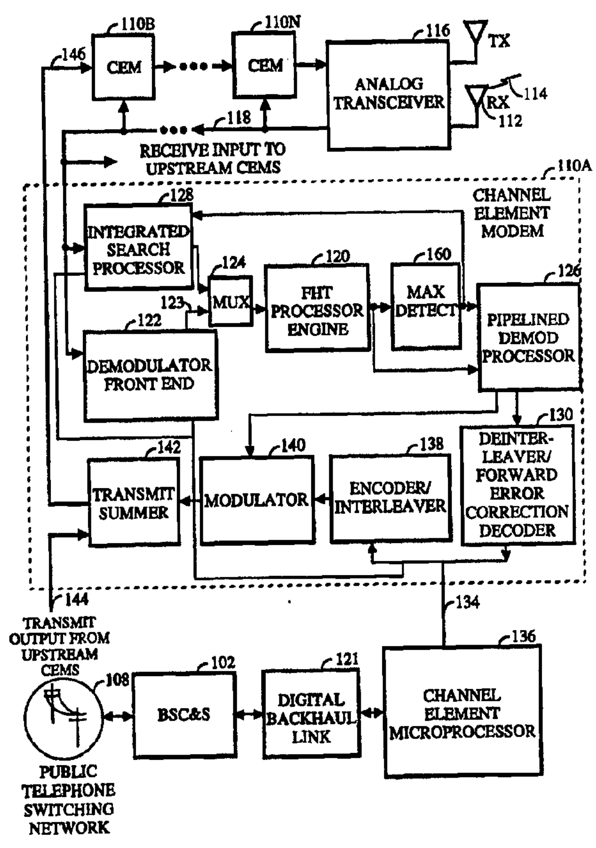
(54) Title: METHOD OF RECEIVING AND SEARCHING A SIGNAL TRANSMITTING IN BURSTS

(57) Abstract

An integrated search processor (128) used in a modem (110) for a spread spectrum communications system buffers in a buffer (172) received signals samples and utilizes a time sliced transform processor (120) operating on successive offsets from the buffer (172). The search processor (128) autonomously steps through a search as configured by a microprocessor (136) specified search parameter set, which can include the group of antennas to search over, the starting offset and width of the search window to search over, and the number of Walsh symbols to accumulate results at each offset. The search processor (128) calculates the correlation energy at each offset, and presents a summary report of the best paths found in the search to use for demodulation element (122) reassignment. The search is done in a linar fashion independent of the probability that a signal being searched for was transmitted at any given time.



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BF	Burkina Faso	**IE**	Ireland	**NZ**	New Zealand
BG	Bulgaria	**IT**	Italy	**PL**	Poland
BJ	Benin	**JP**	Japan	**PT**	Portugal
BR	Brazil	**KE**	Kenya	**RO**	Romania
BY	Belarus	**KG**	Kyrgystan	**RU**	Russian Federation
CA	Canada	**KP**	Democratic People's Republic of Korea	**SD**	Sudan
CF	Central African Republic			**SE**	Sweden
CG	Congo	**KR**	Republic of Korea	**SG**	Singapore
CH	Switzerland	**KZ**	Kazakhstan	**SI**	Slovenia
CI	Côte d'Ivoire	**LI**	Liechtenstein	**SK**	Slovakia
CM	Cameroon	**LK**	Sri Lanka	**SN**	Senegal
CN	China	**LR**	Liberia	**SZ**	Swaziland
CS	Czechoslovakia	**LT**	Lithuania	**TD**	Chad
CZ	Czech Republic	**LU**	Luxembourg	**TG**	Togo
DE	Germany	**LV**	Latvia	**TJ**	Tajikistan
DK	Denmark	**MC**	Monaco	**TT**	Trinidad and Tobago
EE	Estonia	**MD**	Republic of Moldova	**UA**	Ukraine
ES	Spain	**MG**	Madagascar	**UG**	Uganda
FI	Finland	**ML**	Mali	**US**	United States of America
FR	France	**MN**	Mongolia	**UZ**	Uzbekistan
GA	Gabon	**MR**	Mauritania	**VN**	Viet Nam

METHOD OF RECEIVING AND SEARCHING A SIGNAL TRANSMITTED IN BURSTS

BACKGROUND OF THE INVENTION

I. Field of the Invention

The present application is a continuation-in-part application of copending U.S. Patent Application Serial No. 08/316,177, filed September 30, 1994, entitled MULTIPATH SEARCH PROCESSOR FOR A SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM. The present invention relates generally to spread spectrum communication systems and, more particularly, to signal processing in a cellular telephone communication system.

II. Description of the Related Art

In a wireless telephone communication system such as cellular telephone systems, personal communications systems, and wireless local loop system, many users communicate over a wireless channel to connect to wireline telephone systems. Communication over the wireless channel can be one of a variety of multiple access techniques which facilitate a large number of users in a limited frequency spectrum. These multiple access techniques include time division multiple access (TDMA), frequency division multiple access (FDMA), and code division multiple access (CDMA). The CDMA technique has many advantages and an exemplary CDMA system is described in U.S. Patent No. 4,901,307 issued February 13, 1990 to K. Gilhousen et al., entitled "SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM USING SATELLITE OR TERRESTRIAL REPEATERS," assigned to the assignee of the present invention and incorporated herein by reference.

In the just mentioned patent, a multiple access technique is disclosed where a large number of mobile telephone system users, each having a transceiver, communicate through satellite repeaters or terrestrial base stations using CDMA spread spectrum communication signals. In using CDMA communications, the frequency spectrum can be reused multiple times thus permitting an increase in system user capacity.

The CDMA modulation techniques disclosed in U.S. Patent No. 4,901,307 offer many advantages over narrow band modulation techniques used in communication systems using satellite or terrestrial channels. The terrestrial channel poses special problems to any

communication system particularly with respect to multipath signals. The use of CDMA techniques permits the special problems of the terrestrial channel to be overcome by mitigating the adverse effect of multipath, e.g. fading, while also exploiting the advantages thereof.

The CDMA techniques as disclosed in U.S. Patent No. 4,901,307 contemplate the use of coherent modulation and demodulation for both directions of the link in remote unit-satellite communications. Accordingly, disclosed therein is the use of a pilot carrier signal as a coherent phase reference for the satellite-to-remote unit link and the base station-to-remote unit link. In the terrestrial cellular environment, however, the severity of multipath fading with the resulting phase disruption of the channel, as well as the power required to transmit a pilot carrier signal from the remote unit, precludes usage of coherent demodulation techniques for the remote unit-to-base station link. U.S. Patent No. 5,103,459 entitled "SYSTEM AND METHOD FOR GENERATING SIGNAL WAVEFORMS IN A CDMA CELLULAR TELEPHONE SYSTEM", issued June 25, 1990, assigned to the assignee of the present invention, the disclosure of which is incorporated by this reference, provides a means for overcoming the adverse effects of multipath in the remote unit-to-base station link by using noncoherent modulation and demodulation techniques.

In a CDMA cellular telephone system, the same frequency band can be used for communication in all base stations. At the base station receiver, separable multipath, such as a line of site path and another path reflecting off of a building, can be diversity combined for enhanced modem performance. The CDMA waveform properties that provide processing gain are also used to discriminate between signals that occupy the same frequency band. Furthermore, the high speed pseudonoise (PN) modulation allows many different propagation paths of the same signal to be separated, provided the difference in path delays exceeds the PN chip duration. If a PN chip rate of approximately 1 MHz is employed in a CDMA system, the full spread spectrum processing gain, equal to the ratio of the spread bandwidth to the system data rate, can be employed against paths having delays that differ by more than one microsecond. A one microsecond path delay differential corresponds to differential path distance of approximately 300 meters. The urban environment typically provides differential path delays in excess of one microsecond.

The multipath properties of the terrestrial channel produce at the receiver signals having traveled several distinct propagation paths. One characteristic of a multipath channel is the time spread introduced in a signal

that is transmitted through the channel. For example, if an ideal impulse is transmitted over a multipath channel, the received signal appears as a stream of pulses. Another characteristic of the multipath channel is that each path through the channel may cause a different attenuation factor. For example, if an ideal impulse is transmitted over a multipath channel, each pulse of the received stream of pulses generally has a different signal strength than other received pulses. Yet another characteristic of the multipath channel is that each path through the channel may cause a different phase on the signal. For example, if an ideal impulse is transmitted over a multipath channel, each pulse of the received stream of pulses generally has a different phase than other received pulses.

In the radio channel, the multipath is created by reflection of the signal from obstacles in the environment, such as buildings, trees, cars, and people. In general the radio channel is a time varying multipath channel due to the relative motion of the structures that create the multipath. For example, if an ideal impulse is transmitted over the time varying multipath channel, the received stream of pulses would change in time location, attenuation, and phase as a function of the time that the ideal impulse was transmitted.

The multipath characteristic of a channel can result in signal fading. Fading is the result of the phasing characteristics of the multipath channel. A fade occurs when multipath vectors are added destructively, yielding a received signal that is smaller than either individual vector. For example, if a sine wave is transmitted through a multipath channel having two paths where the first path has an attenuation factor of X dB, a time delay of δ with a phase shift of Θ radians, and the second path has an attenuation factor of X dB, a time delay of δ with a phase shift of $\Theta + \pi$ radians, no signal would be received at the output of the channel.

In narrow band modulation systems such as the analog FM modulation employed by conventional radio telephone systems, the existence of multiple paths in the radio channel results in severe multipath fading. As noted above with a wideband CDMA, however, the different paths may be discriminated in the demodulation process. This discrimination not only greatly reduces the severity of multipath fading but provides an advantage to the CDMA system.

Diversity is one approach for mitigating the deleterious effects of fading. It is therefore desirable that some form of diversity be provided which permits a system to reduce fading. Three major types of diversity exist: time diversity, frequency diversity, and space/path diversity.

Time diversity can best be obtained by the use of repetition, time interleaving, and error correction and detection coding which introduce redundancy. A system comprising the present invention may employ each of these techniques as a form of time diversity.

CDMA by its inherent wideband nature offers a form of frequency diversity by spreading the signal energy over a wide bandwidth. Therefore, frequency selective fading affects only a small part of the CDMA signal bandwidth.

Space and path diversity are obtained by providing multiple signal paths through simultaneous links from a remote unit through two or more base stations and by employing two or more spaced apart antenna elements at a single base station. Furthermore, path diversity may be obtained by exploiting the multipath environment through spread spectrum processing by allowing a signal arriving with different propagation delays to be received and processed separately as discussed above. Examples of path diversity are illustrated in U.S. Patent No. 5,101,501 entitled "SOFT HANDOFF IN A CDMA CELLULAR TELEPHONE SYSTEM", issued March 21, 1992 and U.S. Patent No. 5,109,390 entitled "DIVERSITY RECEIVER IN A CDMA CELLULAR TELEPHONE SYSTEM", issued April 28, 1992, both assigned to the assignee of the present invention.

The deleterious effects of fading can be further controlled to a certain extent in a CDMA system by controlling transmitter power. A system for base station and remote unit power control is disclosed in U.S. Patent No. 5,056,109 entitled "METHOD AND APPARATUS FOR CONTROLLING TRANSMISSION POWER IN A CDMA CELLULAR MOBILE TELEPHONE SYSTEM", issued October 8, 1991, also assigned to the assignee of the present invention.

The CDMA techniques as disclosed in U.S. Patent No. 4,901,307 contemplate the use of relatively long PN sequences with each remote unit user being assigned a different PN sequence. The cross-correlation between different PN sequences and the autocorrelation of a PN sequence, for all time shifts other than zero, both have a nearly zero average value which allows the different user signals to be discriminated upon reception. (Autocorrelation and cross-correlation requires logical "0" take on a value of "1" and logical "1" take on a value of "-1" or a similar mapping in order that a zero average value be obtained.)

However, such PN signals are not orthogonal. Although the cross-correlation essentially averages to zero over the entire sequence length, for a short time interval, such as an information bit time, the cross-correlation is a

random variable with a binomial distribution. As such, the signals interfere with each other in much the same way as they would if they were wide bandwidth Gaussian noise at the same power spectral density. Thus the other user signals, or mutual interference noise, ultimately limits the achievable capacity.

It is well known in the art that a set of n orthogonal binary sequences, each of length n, for n any power of 2 can be constructed, see Digital Communications with Space Applications, S.W. Golomb et al., Prentice-Hall, Inc., 1964, pp. 45-64. In fact, orthogonal binary sequence sets are also known for most lengths which are multiples of four and less than two hundred. One class of such sequences that is easy to generate is called the Walsh function, also known as Hadamard matrices.

A Walsh function of order n can be defined recursively as follows:

$$W(n) = \begin{vmatrix} W(n/2)\ ,\ W(n/2) \\ W(n/2)\ ,\ W'(n/2) \end{vmatrix}$$

where W' denotes the logical complement of W, and $W(1) = \begin{vmatrix} 0 \end{vmatrix}$.

Thus,

$$W(2) = \begin{vmatrix} 0, & 0 \\ 0, & 1 \end{vmatrix},$$

$$W(4) = \begin{vmatrix} 0, & 0, & 0, & 0 \\ 0, & 1, & 0, & 1 \\ 0, & 0, & 1, & 1 \\ 0, & 1, & 1, & 0 \end{vmatrix}, \text{ and}$$

$$W(8) = \begin{vmatrix} 0, & 0, & 0, & 0, & 0, & 0, & 0, & 0 \\ 0, & 1, & 0, & 1, & 0, & 1, & 0, & 1 \\ 0, & 0, & 1, & 1, & 0, & 0, & 1, & 1 \\ 0, & 1, & 1, & 0, & 0, & 1, & 1, & 0 \\ 0, & 0, & 0, & 0, & 1, & 1, & 1, & 1 \\ 0, & 1, & 0, & 1, & 1, & 0, & 1, & 0 \\ 0, & 0, & 1, & 1, & 1, & 1, & 0, & 0 \\ 0, & 1, & 1, & 0, & 1, & 0, & 0, & 1 \end{vmatrix}.$$

A Walsh symbol, sequence, or code is one of the rows of a Walsh function matrix. A Walsh function matrix of order n contains n sequences, each of length n Walsh chips. Each Walsh code has a corresponding Walsh index where the Walsh index refers to the number (1 through n) corresponding to the row in which a Walsh code is found. For example, for n=8 Walsh function matrix given above, the all zeroes row corresponds to

Walsh index 1 and the Walsh code 0, 0, 0, 0, 1, 1, 1, 1 corresponds to Walsh index 5.

A Walsh function matrix of order n (as well as other orthogonal functions of length n) has the property that over the interval of n bits, the cross-correlation between all the different sequences within the set is zero. This can be seen by noting that every sequence differs from every other sequence in exactly half of its bits. It should also be noted that there is always one sequence containing all zeroes and that all the other sequences contain half ones and half zeroes. The Walsh symbol which consists all logical zeros instead of half one's and zero's is called the Walsh zero symbol.

On the reverse link channel from the remote unit to the base station, no pilot signal exists to provide a phase reference. Therefore a method is needed to provide a high-quality link on a fading channel having a low Eb/No (energy per bit/noise power density). Walsh function modulation on the reverse link is a simple method of obtaining 64-ary modulation with coherence over the set of six code symbols mapped into the 64 Walsh codes. The characteristics of the terrestrial channel are such that the rate of change of phase is relatively slow. Therefore, by selecting a Walsh code duration which is short compared to the rate of change of phase on the channel, coherent demodulation over the length of one Walsh code is possible.

On the reverse link channel, the Walsh code is determined by the information being transmitted from the remote unit. For example, a three bit information symbol could be mapped into the eight sequences of W(8) given above. An "unmapping" of the Walsh encoded symbols into an estimate of the original information symbols may be accomplished in the receiver by a Fast Hadamard Transform (FHT). A preferred "unmapping" or selection process produces soft decision data which can be provided to a decoder for maximum likelihood decoding.

An FHT is used to perform the "unmapping" process. An FHT correlates the received sequence with each of the possible Walsh sequences. Selection circuitry is employed to select the most likely correlation value, which is scaled and provided as soft decision data.

A spread spectrum receiver of the diversity or "rake" receiver design comprises multiple data receivers to mitigate the effects of fading. Typically each data receiver is assigned to demodulate a signal which has traveled a different path, either through the use of multiple antennas or due to the multipath properties of the channel. In the demodulation of signals modulated according to an orthogonal signaling scheme, each data receiver correlates the received signal with each of the possible mapping values using

an FHT. The FHT outputs of each data receiver are combined and selection circuitry then selects the most likely correlation value based on the largest combined FHT output to produce a demodulated soft decision symbol.

In the system described in the U.S. Patent No. 5,103,459, the call signal begins as a 9600 bit per second information source which is then converted by a rate 1/3 forward error correction encoder to a 28,800 symbols per second output stream. These symbols are grouped 6 at a time to form 4800 Walsh symbols per second, each Walsh symbol selecting one of sixty-four orthogonal Walsh functions that are sixty-four Walsh chips in duration. The Walsh chips are modulated with a user-specific PN sequence generator. The user-specific PN modulated data is then split into two signals, one of which is modulated with an in-phase (I) channel PN sequence and one of which is modulated with a quadrature-phase (Q) channel PN sequence. Both the I channel modulation and the Q channel modulation provide four PN chips per Walsh chip with a 1.2288 MHz PN spreading rate. The I and the Q modulated data are Offset Quadrature Phase Shift Keying (OQPSK) combined for transmission.

In the CDMA cellular system described in the above-referenced U.S. Patent No. 4,901,307, each base station provides coverage to a limited geographic area and links the remote units in its coverage area through a cellular system switch to the public switched telephone network (PSTN). When a remote unit moves to the coverage area of a new base station, the routing of that user's call is transferred to the new base station. The base station-to-remote unit signal transmission path is referred to as the forward link and the remote unit-to-base station signal transmission path is referred to as the reverse link.

As described above, the PN chip interval defines the minimum separation two paths must have in order to be combined. Before the distinct paths can be demodulated, the relative arrival times (or offsets) of the paths in the received signal must first be determined. The channel element modem performs this function by "searching" through a sequence of potential path offsets and measuring the energy received at each potential path offset. If the energy associated with a potential offset exceeds a certain threshold, a signal demodulation element may be assigned to that offset. The signal present at that path offset can then be summed with the contributions of other demodulation elements at their respective offsets. A method and apparatus of demodulation element assignment based on searcher demodulation element energy levels is disclosed in co-pending U.S. Patent Application Serial No. 08/144,902 entitled "DEMODULATION ELEMENT ASSIGNMENT

IN A SYSTEM CAPABLE OF RECEIVING MULTIPLE SIGNALS," filed October 28, 1993, assigned to the assignee of the present invention. Such a diversity or rake receiver provides for a robust digital link, because all paths have to fade together before the combined signal is degraded.

FIG. 1 shows an exemplary set of signals from a single remote unit arriving at the base station. The vertical axis represents the power received on a decibel (dB) scale. The horizontal axis represents the delay in the arrival time of a signal due to multipath delays. The axis (not shown) going into the page represents a segment of time. Each signal spike in the common plane of the page has arrived at a common time but was transmitted by the remote unit at a different time. In a common plane, peaks to the right were transmitted at an earlier time by the remote unit than peaks to the left. For example, the left-most peak spike **2** corresponds to the most recently transmitted signal. Each signal spike **2 - 7** has traveled a different path and therefore exhibits a different time delay and a different amplitude response. The six different signal spikes represented by spikes **2 - 7** are representative of a severe multipath environment. Typical urban environments produce fewer usable paths. The noise floor of the system is represented by the peaks and dips having lower energy levels. The task of a searcher element is to identify the delay as measured by the horizontal axis of signal spikes **2 - 7** for potential demodulation element assignment. The task of the demodulation element is to demodulate a set of the multipath peaks for combination into a single output. It is also the task of the demodulation elements once assigned to a multipath peak to track that peak as it may move in time.

The horizontal axis can also be thought of as having units of PN offset. At any given time, the base station receives a variety of signals from a single remote unit, each of which has traveled a different path and may have a different delay than the others. The remote unit's signal is modulated by a PN sequence. A copy of the PN sequence is also generated at the base station. At the base station, each multipath signal is individually demodulated with a PN sequence code aligned to its timing. The horizontal axis coordinates can be thought of as corresponding to the PN sequence code offset which would be used to demodulate a signal at that coordinate.

Note that each of the multipath peaks varies in amplitude as a function of time as shown by the uneven ridge of each multipath peak. In the limited time shown, there are no major changes in the multipath peaks. Over a more extended time range, multipath peaks disappear and new paths are created as time progresses. The peaks can also slide to earlier or later offsets as the path distances change as the remote unit moves around in the coverage area of the

base station. Each demodulation element tracks small variations in the signal assigned to it. The task of the searching process is to generate a log of the current multipath environment as received by the base station.

In a typical wireless telephone communication system, the remote unit transmitter may employ a vocoding system which encodes voice information in a variable rate format. For example, the data rate may be lowered due to pauses in the voice activity. The lower data rate reduces the level of interference to other users caused by the remote unit transmissions. At the receiver, or otherwise associated with the receiver, a vocoding system is employed for reconstructing the voice information. In addition to voice information, non-voice information alone or a mixture of the two may be transmitted by the remote unit.

A vocoder which is suited for application in this environment is described in copending U.S. patent application Serial No. 08/363,170, entitled "VARIABLE RATE VOCODER," filed December 23, 1994 and assigned to the assignee of the present invention. This vocoder produces from digital samples of the voice information encoded data at four different rates, e.g. approximately 8,000 bits per second (bps), 4,000 bps, 2,000 bps and 1,000 bps, based on voice activity during a 20 millisecond (ms) frame. Each frame of vocoder data is formatted with overhead bits as 9,600 bps, 4,800 bps, 2,400 bps, and 1,200 bps data frames. The highest rate data frame which corresponds to a 9,600 bps frame is referred to as a "full rate" frame; a 4,800 bps data frame is referred to as a "half rate" frame; a 2,400 bps data frame is referred to as a "quarter rate" frame; and a 1,200 bps data frame is referred to as an "eighth rate" frame. In neither the encoding process nor the frame formatting process is rate information included in the data. When the remote unit transmits data at less than full rate, the duty cycle of the remote units transmitted signal is the same as the data rate. For example, at quarter rate a signal is transmitted from the remote unit only one quarter of the time. During the other three quarters time, no signal is transmitted from the remote unit.

The remote unit includes a data burst randomizer. The data burst randomizer determines during which time periods the remote unit transmits and during which time periods it does not transmit given the data rate of the signal to be transmitted, a remote unit specific identifying number, and the time of day. When operating at less than full rate, the data burst randomizer within the remote unit pseudorandomly distributes the active time periods within the transmission burst. A corresponding data burst randomizer is also included in the base station such that the base station can recreate the pseudorandom distribution based on the time of day and the remote unit

specific identifying number but the base station is unaware, a priori, of the data rate of the transmitted signal.

The eighth rate time periods determine a so called worthy group of time periods. A remote unit operating at quarter rate transmits during the worthy group time periods and another set of pseudorandomly distributed periods. A remote unit operating in half rate transmits during the quarter rate time periods and another set of pseudorandomly distributed periods. A remote unit operating in full rate transmits continually. In this way, regardless of the data rate of the transmitted signal, each time period corresponding to the worthy group is sure to correspond to a time when the corresponding remote unit is transmitting a signal. Further details on the data burst randomizer are described in copending U.S. patent application Serial No. 08/291,647, entitled "DATA BURST RANDOMIZER," filed August 16, 1994, and assigned to the assignee of the present invention.

To conserve system resources for actual data of voice transmissions, the remote unit does not transmit the rate for each frame. Therefore, the receiver must determine the rate at which the data was encoded and transmitted based on the received signal so that the receiver associated vocoder can properly reconstruct the voice information. A method of determining the rate at which burst data was encoded without receiving rate information from the transmitter is disclosed in co-pending U.S. Patent Serial No. 08/233,570, entitled "METHOD AND APPARATUS FOR DETERMINING DATA RATE OF TRANSMITTED VARIABLE RATE DATE IN A COMMUNICATIONS RECEIVER" filed April 26, 1994, and assigned to the assignee of the present invention. The method of determining data rate disclosed in the above mentioned patent application is performed after the signal has been received and demodulated therefore the rate information is not available during the searching process.

At the base station, each individual remote unit signal must be identified from the ensemble of call signals received. A system and method for demodulating a remote unit signal received at a base station is described, for example, in U.S. Patent No. 5,103,459. FIG. 2 is a block diagram of the base station equipment described in U.S. Patent No. 5,103,459 for demodulating a reverse link remote unit signal.

A typical prior art base station comprises multiple independent searcher and demodulation elements. The searcher and demodulation elements are controlled by a microprocessor. In this exemplary embodiment, to maintain a high system capacity, each remote unit in the system does not transmit a pilot signal. The lack of a pilot signal on the reverse link increases

the time needed to conduct a survey of all possible time offsets at which a remote unit signal may be received. Typically, a pilot signal is transmitted at a higher power than the traffic bearing signals thus increasing the signal to noise ratio of the received pilot signal as compared to the received traffic channel signals. In contrast, ideally each remote unit transmits a reverse link signal which arrives with a power level equal to the power level received from every other remote unit therefore having a low signal to noise ratio. Also, a pilot channel transmits a known sequence of data. Without the pilot signal, the searching process must examine all possibilities of what data may have been transmitted.

FIG. 2 shows an exemplary embodiment of a prior art base station. The base station of FIG. 2 has one or more antennas **12** receiving CDMA reverse link remote unit signals **14**. Typically, an urban base station's coverage area is split into three sub-regions called sectors. With two antennas per sector, a typical base station has a total of six receive antennas. The received signals are down-converted to baseband by analog receiver **16** that quantizes the received signal I and Q channels and sends these digital values over signal lines **18** to channel element modem **20**. A typical base station comprises multiple channel element modems like channel element modem **20** (not shown in FIG. 2). Each channel element modem **20** supports a single user. In the preferred embodiment, channel element modem **20** comprises four demodulation elements **22** and eight searchers **26**. Microprocessor **34** controls the operation of demodulation elements **22** and searchers **26**. The user PN code in each demodulation element **22** and searcher **26** is set to that of the remote unit assigned to that channel element modem **20**. Microprocessor **34** steps searchers **26** through a set of offsets, called a search window, that is likely to contain multipath signal peaks suitable for assignment of demodulation elements **22**. For each offset, searcher **26** reports the energy it finds at that offset to microprocessor **34**. Demodulation elements **22** are then assigned by microprocessor **34** to the paths identified by searchers **26**. Once one of demodulation elements **22** has locked onto the signal at its assigned offset, it then tracks that path on its own without supervision from microprocessor **34**, until the path fades away or until it is assigned to a new path by microprocessor **34**.

For the system of FIG. 2, each demodulation element **22** and searcher **26** contains one FHT processor **52** capable of performing one FHT transform during a time period equal to the period of a Walsh symbol. The FHT processor is slaved to "real time" in the sense that every Walsh symbol interval one value is input and one symbol value is output from the FHT.

Therefore, to provide a rapid searching process, more than one searcher **26** must be used. Each of searchers **26** supplies back to microprocessor **34** the results of the search it performs. Microprocessor **34** tabulates these results for use in the assignment of demodulation elements **22** to the incoming signals.

In FIG. 2, the internal structure of only one demodulation element **22** is shown, but should be understood to apply to searchers **26** as well. Each demodulation element **22** or searcher **26** of the channel element modem has a corresponding I PN and Q PN sequence generator **36, 38** and the user-specific PN sequence generator **40** that is used to select a particular remote unit. User-specific PN sequence output **40** is XOR'd by XOR gates **42** and **44** with the output of I PN and Q PN sequence generators **36** and **38** to produce PN-I' and PN-Q' sequences that are provided to despreader **46**. The timing reference of PN generators **36, 38, 40** is adjusted to the offset of the assigned signal, so that despreader **46** correlates the received I and Q channel antenna samples with the PN-I' and PN-Q' sequence consistent with the assigned signal offset. Four of the despreader outputs, corresponding to the four PN chips per Walsh chip, are summed to form a single Walsh chip by accumulators **48** and **50**. The accumulated Walsh chip is then input into Fast Hadamard Transform (FHT) processor **52**. When 64 chips corresponding to one Walsh symbol have been received, FHT processor **52** correlates the set of sixty-four Walsh chips with each of the sixty-four possible transmitted Walsh symbols and outputs a sixty-four entry matrix of soft decision data. The output of FHT processor **52** is then combined with those of other assigned demodulation elements by combiner **28**. The output of combiner **28** is a "soft decision" demodulated symbol, weighted by the confidence that it correctly identifies the originally transmitted Walsh symbol. The soft decision data is then passed to forward error correction decoder **29** for further processing to recover the original call signal. This call signal is then sent through digital link **30**, such as a T1 or E1 link, that routes the call to public switched telephone network (PSTN) **32**.

Like each demodulation element **22**, each searcher **26** contains a demodulation data path with an FHT processor capable of performing one FHT transform during a time period equal to the period of a Walsh symbol. Searcher **26** only differs from demodulation element **22** in how its output is used and in that it does not provide time tracking. For each offset processed, each searcher **26** finds the correlation energy at that offset by despreading the antenna samples, accumulating them into Walsh chips that are input to the FHT transform, performing the FHT transform and summing the maximum FHT output energy for each of the Walsh symbols for which the searcher dwells at an offset. The final sum is reported back to microprocessor **34**.

Generally each searcher **26** is stepped through the search window with the others as a group by microprocessor **34**, each separated from its neighbor by half of a PN chip. In this way enough correlation energy exists at each maximum possible offset error of a quarter chip to ensure that a path is not missed because the searcher did not correlate with the exact offset of the path. After sequencing searchers **26** through the search window, microprocessor **34** evaluates the results reported back, looking for strong paths for demodulation elements assignment as described in above mentioned co-pending U.S. Patent Application Serial No. 08/144,902.

The multipath environment is constantly changing as the remote unit and other reflective objects move about in the base station coverage area. The number of searches that must be performed is set by the need to find multipath quickly enough so that valid paths may be put to good use by the demodulation elements. On the other hand, the number of demodulation elements required is a function of the number of paths generally found to be usable at any point in time. To meet these needs, the system of FIG. 2 has two searchers **26** and one demodulation element **22** for each of four demodulator integrated circuits (IC's) used, for a total of four demodulation elements and eight searchers per channel element modem. Each of these twelve processing elements contains a complete demodulation data path, including the FHT processor which takes a relatively large, costly amount of area to implement on an integrated circuit. In addition to the four demodulator IC's the channel element modem also has a modulator IC and a forward error correction decoder IC for a total of 6 IC chips. A powerful and expensive microprocessor is needed to manage and coordinate the demodulation elements and the searchers. As implemented in FIG. 2, these circuits are completely independent and require the close guidance of microprocessor **34** to sequence through the correct offsets, and handle the FHT outputs. Every Walsh symbol microprocessor **34** receives an interrupt to process the FHT outputs. This interrupt rate alone necessitates use of a high powered microprocessor.

It would be advantageous if the six IC's required for a modem could be reduced to a single IC needing less microprocessor support, thereby reducing the direct IC cost and board-level production cost of the modem, and allowing migration to a lower cost microprocessor (or alternately a single high powered microprocessor supporting several channel element modems at once.) Just relying on shrinking feature sizes of the IC fabrication process and placing the six chips together on a single die is not enough. The fundamental architecture of the searcher needs to be redesigned for a truly cost effective single chip modem. From the discussion above, it should be apparent that

there is a need for a signal receiving and processing apparatus that can demodulate a spread spectrum call signal at a lower cost and in a more architecturally efficient manner.

The present invention can use a set of real time searchers as described above or a single, integrated search processor that can quickly evaluate large numbers of offsets that potentially contain multipath of a received call signal.

The present invention is a method of searching for a multipath signal which is transmitted at an unknown variable rate and is subjected to power control.

SUMMARY OF THE INVENTION

The present invention is a method of searching for a multipath signal which is transmitted at an unknown variable rate and is subjected to power control. The searching method is linear in that no attempt is made to synchronize the searching process to time known to contain data. The searching process is aligned to power control group boundaries so that accurate power estimates can be made.

BRIEF DESCRIPTION OF THE DRAWINGS

The features, objects, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1 represents an exemplary severe multipath signal condition;

FIG. 2 is a block diagram of a prior art communications network demodulation system;

FIG. 3 represents an exemplary CDMA telecommunications system constructed in accordance with the present invention;

FIG. 4 is a block diagram of a channel element modem constructed in accordance with the present invention;

FIG. 5 is a block diagram of the search processor;

FIG. 6 illustrates the circular nature of the antenna sample buffer using a first offset;

FIG. 7 illustrates the circular nature of the antenna sample buffer for a second accumulation at the first offset of FIG. 6;

FIG. 8 illustrates the circular nature of the antenna sample buffer for a second offset;

FIG. 9 is a graph showing how the searcher processes the receiver input as a function of time;

FIG. 10 is a block diagram of the searcher front end;

FIG. 11 is a block diagram of the searcher despreader;

FIG. 12 is a block diagram of the searcher result processor;

FIG. 13 is a block diagram of the searcher sequencing control logic;

FIG. 14 is a timing diagram showing the processing sequence depicted in FIG. 5, showing the corresponding states of certain control logic elements presented in FIG. 13; and

FIG. 15 is an alternative block diagram of the search processor.

DESCRIPTION OF THE PREFERRED EMBODIMENT

In the following description of a method and system for processing telephone calls within a digital wireless telephone system, various references are made to processes and steps that are performed in order to achieve a desired result. It should be understood that such references do not describe human actions or thought, but are directed towards the operation, modification and transformation of various systems including and especially those systems which process electrical and electromagnetic signals and charges, optical signals, or a combination thereof. Fundamental to such systems is the use of various information storage devices, often referred to as "memory", which store information via the placement and organization of atomic or super-atomic charged particles on hard disk media or within silicon, gallium arsenic, or other semiconductor based integrated circuit media, and the use of various information processing devices, often referred to as "microprocessors," which alter their condition and state in response to such electrical and electromagnetic signals and charges. Memory and microprocessors that store and process light energy or particles having special optical characteristic, or a combination thereof, are also contemplated and use thereof is consistent with the operation of the described invention.

The present invention can be implemented in a wide variety of data transmission applications and in the preferred embodiment illustrated in FIG. 3 is implemented within system **100** for voice and data transmission in which system controller and switch (BSC&S) **102** performs interface and control functions to permit call communications with remote units **104** through base stations **106**. BSC&S **102** controls the routing of calls between public switched telephone network (PSTN) **108** and base stations **106** for transmission to and from remote units **104**.

FIG. 4 illustrates channel element modems **110A - 110N** and other elements of the base station infrastructure operating in accordance with the CDMA methods and data formats described in the above-referenced patents.

A plurality of antennas **112** provides received reverse link signal **114** to analog transmitter receiver (transceiver) **116**. Analog transceiver **116** down-converts reverse link signal **114** to baseband and samples the baseband waveform at eight times the PN chip rate of the CDMA received signal as defined above. Analog transceiver **116** provides the digital antenna samples to channel element modems **110A - 110N** through base station RX backplane signal **118**. Each channel element modem **110A - 110N** may be assigned to one remote unit having an active communication established with the base station. Each channel element modem **110A - 110N** is nearly identical in structure.

When channel element modem **110A** is assigned to an active call, demodulator front end **122** and integrated search processor **128** isolate a signal from the corresponding remote unit from the plurality of call signals contained in reverse link signal **114** by use of the PN sequences as described in the above referenced patents and patent applications. Channel element modem **110A** includes single integrated search processor **128** to identify multipath signals that can be used by demodulator front end **122**. In the preferred embodiment, time sliced FHT processor engine **120** services both integrated search processor **128** and demodulator front end **122**. Other than sharing FHT processor engine **120** and related max detect block **160**, integrated search processor **128**, is stand-alone, self-controlled, and self-contained. A searching architecture is detailed in a co-pending U.S. Patent Application Serial No. 08/316,177 entitled "MULTIPATH SEARCH PROCESSOR FOR A SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM" filed September 30, 1994, and assigned to the assignee of the present invention.

FHT processor engine **120** is the core of the demodulation process. In the preferred embodiment, FHT processor engine **120** correlates the received Walsh symbol values with each of the possible Walsh symbols that may have been transmitted by the remote unit. FHT processor engine **120** outputs a correlation energy corresponding to each possible Walsh symbol where a higher correlation energy level corresponds to a higher likelihood that the symbol corresponding to that Walsh index was communicated by the remote unit. Max detect block **160** then determines the largest of the 64 FHT transform energy outputs. The maximum correlation energy and the corresponding Walsh index from max detect block **160** and each of the 64 correlation energy output from FHT processor engine **120** are passed to pipelined demodulator processor **126** for further signal processing. The maximum correlation energy and the corresponding Walsh index from max detect block **160** are passed back to integrated search processor **128**.

Pipelined demodulator processor **126** time aligns and combines symbol data received at different offsets into a single demodulated "soft decision" symbol stream. In addition, pipelined demodulator processor **126** calculates the power level of the signal being received. From the received power level a power control indication is created to command the remote unit to raise or lower the remote unit's transmit power. The power control indication is passed through modulator **140** which adds the indication to the base station transmitted signal for reception by the remote unit. This power control loop operates under the method described in U.S. Patent 5,056,109 referenced above.

The soft decision symbol stream from pipelined demodulator processor **126** is output to deinterleaver/forward error correction decoder **130** where it is deinterleaved and decoded. Channel element microprocessor **136** supervises the entire demodulation procedure and obtains the recovered data from deinterleaver/forward error correction decoder **130** via microprocessor bus interface **134**. The data is then routed through digital backhaul link **121** to BSC&S **102** which connects the call through PSTN **108**.

The forward link data path proceeds much as the inverse of the functions just presented for the reverse link. The signal is provided from PSTN **108** through BSC&S **102** and to digital backhaul **121**. Digital backhaul **121** provides input to encoder/interleaver **138** through channel element microprocessor **136**. After encoding and interleaving the data, encoder/interleaver **138** passes the data to modulator **140** where it is modulated as disclosed in the above referenced patents. Output **146** of modulator **140** is passed to transmit summer **142** where it is added to the outputs of other channel element modems **110B - 110N** prior to being up converted from baseband and amplified in analog transmitter receiver **116**. A summing method is disclosed in a co-pending U.S. Patent Application Serial No. 08/316,156 entitled "SERIAL LINKED INTERCONNECT FOR THE SUMMATION OF MULTIPLE DIGITAL WAVEFORMS," filed September 30, 1994, and assigned to the assignee of the present invention. As presented in the above referenced patent application, the transmit summer corresponding to each of channel element modems **110A - 110N** can be cascaded in a daisy-chain fashion eventually resulting in a final sum that is provided to analog transceiver **116** for broadcasting.

FIG. 5 shows the elements comprising integrated search processor **128**. The heart of the searching process is time sliced FHT processor engine **120**, which, as mentioned above, is shared between integrated search processor **128** and demod front end **122** (not shown in FIG. 5). FHT processor engine **120** can

perform Walsh symbol transforms at a rate 32 times faster than FHT processor **52** of FIG. 2. This rapid transform capability empowers the time sliced operation of channel element modem **110**.

In the preferred embodiment FHT processor engine **120** is constructed using a six stage butterfly network. As explained in detail above, a Walsh function of order n can be defined recursively as follows:

$$W(n) = \begin{vmatrix} W(n/2)\ ,\ W(n/2) \\ W(n/2)\ ,\ W'(n/2) \end{vmatrix}$$

where W' denotes the logical complement of W, and W(1) = 0.

In the preferred embodiment a Walsh sequence is generated where n = 6, therefore a 6-stage butterfly trellis is used to correlate the 64 Walsh chips of one transmitted Walsh symbol with each of the 64 possible Walsh sequences. A structure and method of operation for FHT processor engine **120** is detailed in a co-pending U.S. Patent Application Serial No. 08/173,460 entitled "METHOD AND APPARATUS FOR PERFORMING A FAST HADAMARD TRANSFORM," filed December 22, 1993, assigned to the assignee of the present invention.

To reap the benefits of FHT processor engine **120** with thirty-two times the throughput of its real-time-slaved counterpart, FHT processor engine **120** must be provided with high rate input data to process. Antenna sample buffer **172** has been specifically tailored to meet this need. Antenna sample buffer **172** is written to and read from in a circular manner

The searching process is grouped in sets of single offset searches. The highest level of grouping is the antenna search set. Each antenna search set is made up of a plurality of search windows. Typically each search window in the antenna search set is an identically performed search group where each search window in the antenna search receives data from a different antenna. Each search window is made up of a series of search rakes. A search rake is a set of sequential search offsets that is performed in a time equivalent to the duration of a Walsh symbol. Each search rake is comprised of a set of rake elements. Each rake element represents a single search at a given offset.

At the beginning of the searching process, channel element microprocessor **136** sends parameters specifying a search window which may be part of an antenna search set. The width of the search window may be designated in PN chips. The number of search rakes needed to complete the search window varies depending on the number of PN chips specified in the

search window. The number of rake elements per search rake can be specified by channel element microprocessor **136** or could be fixed to some constant.

Referring again to FIG. 1 showing an exemplary set of signals arriving at the base station from a single remote unit, the relationship of the search window, search rake, and rake element becomes more clear. The vertical axis in FIG. 1 represents the power received in decibels (dB). The horizontal axis represents the delay in the arrival time of a signal due to multipath delays. The axis (not shown) going into the page represents a segment of time. Each signal spike in the common plane of the page has arrived at the same time but has been transmitted by the remote unit at a different time.

The horizontal axis can be thought of as having units of PN chip offset. At any given time, the base station perceives a variety of signals from a single remote unit, each of which has traveled a different path and may have a different delay than the others. The remote unit's signal is modulated by a PN sequence. A copy of the PN sequence is also generated at the base station. At the base station, if each multipath signal were individually demodulated, a PN sequence code aligned to each signal's timing would be needed. Each of these aligned PN sequences would be delayed from the zero offset reference at the base station due to the delay. The number of PN chips that the aligned PN sequence is delayed from the zero offset base station reference could be mapped to the horizontal axis.

In FIG. 1, time segment **10** represents a search window set of PN chip offsets to be processed. Time segment **10** is divided into five different search rakes such as search rake time segment **9**. Each search rake is in turn made up of a number of rake elements which represent the actual offsets to be searched. For example, in FIG. 1, each search rake is made up of 8 different rake elements such as the rake element indicated by arrow **8**.

To process a single rake element as indicated by arrow **8**, a set of samples over time at that offset are needed. For example, to process the rake element indicated by arrow **8**, the despreading process needs the set of sample at the offset indicated by arrow **8** going back into the page over time. The despreading process also needs a corresponding PN sequence. The PN sequence can be determined by noting the time the samples arrived and the offset desired to be processed. The desired offset can be combined with the arrival time to determine the corresponding PN sequence to be correlated with the received samples.

As the rake element is despread the receive antenna samples and the PN sequence are run through a series of values over time. Note that the received antenna samples are the same for all offsets shown in FIG. 1 and

spikes **2-7** are showing exemplary multipath peaks which arrive simultaneously and are only discriminated by the despreading process.

In the preferred embodiment described below, each rake element is offset in time from the preceding rake element by one half PN chip in time. This means that if the rake element corresponding to arrow **8** was correlated beginning from the sliced plane shown and moving forward in time (into the page as shown) then the rake element to the left of the one corresponding to arrow **8** would use samples starting one half chip in time back from the sliced plane shown. This progression in time allows each rake element in a common search rake to be correlated to the same PN sequence.

Each remote unit receives the base station's transmitted signal delayed by some amount due to the path delay through the terrestrial environment. The same I and Q PN short code and the user PN long code generation is also being performed in the remote unit. The remote unit generates a time reference based on the time reference it perceives from the base station. The remote unit uses the time reference signal as an input to its I and Q PN short code and the user PN long code generators. The information signal received at the base station from the remote unit is therefore delayed by the round trip delay of the signal path between the base station and the remote unit. Therefore if the timing of the PN generator used in the searching process is slaved to the zero offset timing reference at the base station, the output of the generators is always available before the corresponding signal is received from the remote unit.

In an OQPSK signal, the I channel data and the Q channel data are offset from each other by one half chip in time. Therefore OQPSK despreading used in the preferred embodiment requires data sampled at twice the chip rate. The searching process also operates optimally with data sampled at half the chip rate. Each rake element within a search rake is offset by one half chip from the previous rake element. The one half chip rake element resolution ensures that multipath peak signals are not skipped over without detection. For these reasons antenna sample buffer **172** of FIG. 5 stores data sampled at twice the PN chip rate.

One Walsh symbol worth of data is read from antenna sample buffer **172** to process a single rake element. For each successive rake element, one Walsh symbol worth of data is read out of antenna sample buffer **172** one half of a PN chip offset from the previous rake element. Each rake element is despread with the same PN sequence read from PN sequence buffer **176** by despreader **178** for each rake element in the search rake.

Antenna sample buffer **172** is two Walsh symbols deep and is continually and repeatedly read from and written to throughout the searching process. Within each search rake, the rake element having the latest offset in time is processed first. The latest offset corresponds to the signal which has traveled the longest signal path from the remote unit to the base station. The time at which the searcher starts to process a search rake is keyed to the Walsh symbol boundaries associated with the rake element having the latest offset in the search rake. A time strobe, referred to as the offset Walsh symbol boundary, indicates the earliest time that all of the samples needed are available in antenna sample buffer **172** and the searching process can begin the first rake element in the search rake.

The operation of antenna sample buffer **172** is most easily illustrated by noting its circular nature. FIG. 6 shows an illustrative diagram of the operation of antenna sample buffer **172**. In FIG. 6 thick circle **400** can be thought of as antenna sample buffer **172** itself. Antenna sample buffer **172** contains memory locations for two Walsh symbols worth of data. Write pointer **406** circulates around antenna sample buffer **172** in the direction indicated in real time, meaning that write pointer **406** rotates around the two Walsh symbol deep antenna sample buffer **172** in the time that two Walsh symbols worth of samples are passed to searcher front end **174**. As the samples are written into antenna sample buffer **172** according to the memory location indicated by write pointer **406**, the previously stored values are overwritten. In the preferred embodiment, antenna sample buffer **172** contains 1024 antenna samples because each of the two Walsh symbols contains 64 Walsh chips, each Walsh chip contains 4 PN chips, and each PN chip is sampled twice.

The operation of the searching process is divided into discrete 'time slices.' In the preferred embodiment, a time slice is equal to 1/32 of the Walsh symbol duration. The choice of 32 time slices per Walsh symbol is derived from the available clocking frequency and number of clock cycles need to perform an FHT. 64 clock cycles are required to perform an FHT for one Walsh symbol. In the preferred embodiment, a clock running at eight times the PN chip frequency is available and provides the necessary performance level. Eight times the PN chip rate multiplied by the 64 required clocks is equivalent to the time it takes to receive two Walsh chips worth of data. Because there are 64 Walsh chips in each half of the buffer, 32 time slices are needed to read in a complete Walsh symbol.

In FIG. 6, a set of concentric arcs outside of thick circle **400** represents read and write operation with antenna sample buffer 172. (The arcs within

thick circle **400** are used to aid explanation and do not correspond to read or write operations.) Each arc represents a read or write operation during one time slice. The arc closest to the center of the circle occurs first in time and each successive arc represents an operation occurring in successively later time slices as indicated by time arrow **414**. Each of the concentric arcs corresponds to a section of antenna sample buffer **172** as represented by thick circle **400**. If one were to imagine radii drawn from the center of thick circle **400** to the end points of each of the concentric arcs, the portion of thick circle **400** between the intersection of the radii and thick circle **400** would be representative of the memory locations accessed. For example, during the first time slice operation shown, 16 antenna samples are written to antenna sample buffer **172** represented by arc **402A**.

In FIGS 6, 7, and 8 the following search parameters for the illustrative search window are assumed:

Search window width = 24 PN chips
Search offset = 24 PN chips
Number of symbols to accumulate = 2
Number of rake elements per search rake =24.

FIG. 6 also assumes that antenna sample buffer **172** contains nearly a full Walsh symbol worth of valid data before the write indicated by arc **402A**. During subsequent time slices, a write corresponding to arc **402B** and to arc **402C** occurs. During the 32 time slices available during one Walsh symbol worth of time, the write operations continue from arc **402A** to arc **402FF** most of which are not shown.

The 32 time slices represented by arcs **402A** to **402FF** correspond to the time used to complete one search rake. Using the parameters given above, the search rake begins 24 PN chips offset from zero offset reference or 'real time' and contains 24 rake elements. The 24 PN chip offset corresponds to a rotation 16.875 degrees around thick circle **400** from the beginning of the first write indicated by arc **402A** (calculated by dividing the 24 PN chip offset by the 256 total number of chips in half antenna sample buffer **172** and multiplying by 180 degrees.) The 16.875 degree arc is illustrated by arc **412**. The 24 rake elements correspond to reads indicated by arcs **404A** - **404X** most of which are not shown. The first read corresponding to arc **404A** begins at the search offset some time after the write corresponding to **402C** so that a contiguous set of data is available. Each successive read such as **404B** is offset from the previous by a single memory location, corresponding to a 1/2 PN chip of time. During the search rake shown, the reads move toward earlier time offsets as shown by arcs **404A** - **404X** slanting counter clockwise with

progressing time in the opposite rotation direction as write pointer indication **406**. The 24 read represented by arcs **404A** to **404X** traverse the arc indicated by arc **418**. The progression of the reads toward earlier samples has the advantage of providing seamless searching within a search window as each search rake is executed. This advantage is explained in detail subsequently herein.

Each of the reads corresponding to arcs **404A** to **404X** passes one Walsh symbol worth of data to despreader **178**. The read therefore corresponds to traversing thick circle **400** by 180 degrees. Note that in the search rake shown in FIG. 6, the last write corresponding to arc **402FF**, and last read corresponding to arc **404X** do not include any common memory locations to ensure contiguous valid data. However, hypothetically, if the pattern of read and writes were to continue they would in fact intersect and valid data would not be provided under this condition.

In most signaling conditions, the result of a rake element worth of data collected during one Walsh symbol worth of time is not sufficient to provide accurate information about the location of diverse signals. In these cases, a search rake may be repeated multiple times. The results of rake elements in successive search rakes at a common offset are accumulated by search result processor **162** as explained in detail subsequently herein. In this case the search parameters given above indicate that the number of symbols to accumulate at each offset is two. FIG. 7 shows the search rake of FIG. 6 repeated at the same offset for the next successive Walsh symbol worth of data. Note that antenna sample buffer **172** contains two Walsh symbols worth of data so that the data that is needed for processing during the search rake indicated on FIG. 7 was written during the search rake shown on FIG. 6. In this configuration, memory locations 180 degrees away from each other represent the same PN offset.

After completing the two accumulated search rakes in FIGS 6 and 7, the searching process advances to the next offset in the search window. The amount of the advance is equal to the width of the search rake processed, in this case 12 PN chips. As specified in the search parameters, the search window width is 24 PN chips. The width of the window will determine how many search rake offsets are needed to complete the search window. In this case two different offsets are needed to cover the 24 PN chip window width. The window width is indicated on FIG. 8 by arc **412**. The second offset for this search window begins at the offset following the last offset of the previous search rake and continues around to the nominal zero offset point as set by the location of the beginning of the first write as indicated by arc **430A**. Again

there are 24 rake elements within the search rake as indicated by arcs **432A - 432X** most of which are not shown. Again the 32 writes are indicated by the arcs **430A - 430FF**. Thus the last write, as indicated by arc **430FF**, and the last read, as indicated by arc **432X**, abut one another in antenna sample buffer **172**.

The search rake shown in FIG. 8 is repeated on the opposite side of antenna sample buffer **172** much as the search rake in FIG. 6 is repeated in FIG. 7 because the search parameters designate that each symbol is accumulated twice. At the completion of the second accumulation of the second search rake, integrated search processor **128** is available to begin another search window. The subsequent search window could have a new offset or it could specify a new antenna or both.

In FIG. 8, the location of the boundary between the read half and the write half of the buffer is marked with label **436**. In FIG. 6, the boundary is marked with label **410**. The signal which indicates the point in time corresponding to labels **410** and **436** is referred to as the offset Walsh symbol strobe and also indicates that a new Walsh symbol worth of samples is available. As the search rakes within a window advance to earlier offsets, the boundary between the read and write halves of the buffer slews in lock step counterclockwise as shown in FIG. 8. If after the completion of the present search window, if a large change in the offset being processed is desired, the offset Walsh symbol strobe may be advanced a large portion of the circumference of the circle.

FIG. 9 is a search timeline that provides further graphical illustration of the searcher processing. Time is plotted along the horizontal axis in units of Walsh symbols. Antenna sample buffer **172** addresses and PN sequence buffer **176** addresses are shown along the vertical axis, also in units of Walsh symbols. Because antenna sample buffer **172** is two Walsh symbols deep, antenna sample buffer **172** addressing wraps on even Walsh symbol boundaries, but for illustrative purposes, FIG. 9 shows the addresses before being folded on top of one another. Samples are written into antenna sample buffer **172** at an address taken directly from the time they were obtained, so write pointer **181** into antenna sample buffer **172** is a straight forty-five degree inclined line. The offset being processed maps into a base address in antenna sample buffer **174** to start a read of a Walsh symbol of samples for a single rake element. The rake elements are illustrated in FIG. 9 as nearly vertical read pointer line segments **192**. Each rake element maps to a Walsh symbol in height as referred to the vertical axis and 1/32 of a Walsh symbol as referred to the horizontal axis.

The vertical gaps between the rake elements within a search rake are caused by demod front end **122** interrupting the search process to use FHT processor engine **120**. Demod front end **122** operates in real time and has first priority use of FHT processor engine **120** whenever it has a current or queued set of data for processing. Therefore typically use of FHT processor engine **120** is given to demod front end **120** on each Walsh symbol boundary corresponding to a PN offset that is being demodulated by demod front end **122**.

FIG. 9 shows the same search rakes shown in FIGS. 6, 7, and 8. For example, search rake **194** has 24 rake elements each of which corresponds to one to the read arcs **404A - 404X** on FIG. 6. On FIG. 9 for search rake **194**, pointer **410** indicates the offset Walsh symbol strobe corresponds to the like pointer on FIG. 6. To read the current samples, each rake element must be beneath write pointer **181**. The downward slope of the rake elements with a search rake indicates the steps towards earlier samples. Search rake **195** corresponds to the search rake shown in FIG. 7 and search rake **196** corresponds to the search rake shown in FIG. 8.

In the search window defined by the parameters above, only 24 rake elements per search rake are specified even though the search rake has 32 available time slices. Each rake element can be processed in one time slice. However, it is not practically possible to increase the number of rake elements per search rake to 32 to match the number of time slices available during a search rake. Demod front end **122** uses some of the available time slices of the FHT processor. There is also a time delay associated with a rake advance as the read process must wait for the write process to fill the buffer with valid data at the advanced offset. Also some margin is needed to synchronize to a time slice processing boundary after observing the offset Walsh symbol strobe. All these factors practically limit the number of rake elements which can be processed in a single search rake. In some cases the number of rake elements per search rake could be increased such as if demod front end **122** has only one demodulation element assigned and hence only interrupts FHT processor engine **120** once per search rake. Therefore in the preferred embodiment, the number of rake elements per search rake is controllable by channel element microprocessor **136**. In alternative embodiments, the number of rake elements per search rake could be a fixed constant.

There also can be significant overhead delay when switching between source antennas at the input to the sample buffer or changing the search window starting point or width between searches. If one rake needs a particular set of samples and the next rake for a different antenna needs to use

an overlapping part of the buffer, the next rake must postpone processing until the next offset Walsh symbol boundary occurs, at which point a complete Walsh symbol of samples for the new antenna source is available. In FIG. 9, search rake **198** is processing data from a different antenna than search rake **197**. Horizontal line **188** indicates the memory location corresponding to the new antenna input samples. Note that search rake **197** and **198** do not use any common memory locations.

For every time slice, two Walsh chips of samples must be written to the sample buffer and one full Walsh symbol of samples may be read from the sample buffer. In the preferred embodiment, there are 64 clock cycles during each time slice. A full Walsh chip of samples is comprised of four sets of samples: ontime I channel samples, late I channel samples, ontime Q channel samples and late Q channel samples. In the preferred embodiment, each sample is four bits. Therefore sixty four bits per clock are needed from antenna sample buffer **172**. Using a single port RAM, the most straightforward buffer design doubles the word width to 128 bits, and splits the buffer into two 64 bit wide, 64 word, independently read/writeable even and odd Walsh chip buffers **168, 170**. The much less frequent writes to the buffer are then multiplexed in between reads, which toggle between the two banks on successive clock cycles.

The Walsh chip samples read from the even and odd Walsh chip buffers **168, 170** has an arbitrary alignment to the physical RAM word alignment. Therefore on the first read of a time slice, both halves are read into despreader **178** to form a two Walsh chip wide window from which the single Walsh chip with the current offset alignment is obtained. For even Walsh chip search offsets, the even and odd Walsh chip buffer address for the first read are the same. For odd Walsh chip offsets, the even address for the first read is advanced by one from the odd address to provide a consecutive Walsh chip starting from the odd half of the sample buffer. The additional Walsh chips needed by despreader **178** can be passed thereto by a read from a single Walsh chip buffer. Successive reads then ensure that there is always a refreshed two Walsh chip wide window from which to draw a Walsh chip of data aligned to the current offset being processed.

Referring again to FIG. 5, for each rake element in a search rake, the same Walsh symbol of PN sequence data from PN sequence buffer **176** is used in the despreading process. For every clock cycle of a time slice, four pairs of PN-I' and PN-Q' are needed. Using a single port RAM, the word width is doubled and read from half as often. The single write to PN sequence

buffer **176** needed per time slice is then performed on a cycle not used for reading.

Because the searching process can specify searching PN offsets of up to two Walsh symbols delayed from the current time, four Walsh symbols worth of PN sequence data must be stored. In the preferred embodiment PN sequence buffer **176** is a one hundred and twenty eight word by sixteen bit RAM. Four Walsh symbols are required because the starting offset can vary by 2 Walsh symbols and once the starting offset is chosen, one Walsh symbol worth of PN sequence is need for correlation meaning three Walsh symbols worth of data is need for the despreading process. Because the same PN sequence is repeatedly used, the data in PN sequence buffer **176** cannot be overwritten during the despreading process corresponding to a single search rake. Therefore an additional Walsh symbol worth of memory is needed to store the PN sequence data as it is generated.

The data that is written into both PN sequence buffer **176** and antenna sample buffer **172** is provided by searcher front end **174**. A block diagram of searcher front end **174** is shown in FIG. 10. Searcher front end **174** includes short code I and Q PN generators **202, 206** and the long code User PN generator **204**. The values output by short code I and Q PN generators **202, 206** and the long code User PN generator **204** are determined in part by the time of day. Each base station has a universal timing standard such as GPS timing to create a timing signal. Each base station also transmits its timing signal over the air to the remote units. At the base station, the timing reference is said to have zero offset because it is aligned to the universal reference.

The output of long code User PN generator **204** is logically XOR'd with the output of short code I and Q PN generators **202, 206** by XOR gates **208** and **210** respectively. (This same process is also performed in the remote unit and the output is used to modulate the remote unit's transmitted signal.) The output of XOR gates **208** and **210** is stored in serial to parallel shift register **212**. Serial to parallel shift register **212** buffers the sequences up to the width of PN sequence buffer **176**. The output of serial to parallel shift register **212** is then written into PN sequence buffer **176** at an address taken from the zero offset reference time. In this way, searcher front end **174** provides the PN sequence data to PN sequence buffer **176**.

Searcher front end **174** also provides antenna samples to antenna sample buffer **172**. Receive samples **118** are selected from one of a plurality of antennas via a MUX **216**. The selected receive samples from MUX **216** are passed to latch **218** where they are decimated, meaning one quarter of the samples are selected for use in the searching process. Receive samples **118**

have been sampled at eight times the PN chip rate by analog transmitter receiver **116** (of FIG. 4). Processing within the searching algorithm is designed for samples taken at one half the chip rate. Therefore only one quarter of the received samples need be passed to antenna sample buffer **172**.

The output of the latch **218** is fed to serial to parallel shift register **214**, which buffers the samples up to the width of antenna sample buffer **172**. The samples are then written into even and odd Walsh chip buffers **168, 170** at addresses also taken from the zero offset reference time. In this way, despreader **178** can align the antenna sample data with a known offset with respect to the PN sequence.

Referring again to FIG. 5, for each clock cycle in a time slice, despreader **178** takes a Walsh chip of antenna samples from antenna sample buffer **172** and a corresponding set of PN sequence values from PN sequence buffer **176** and outputs an I and Q channel Walsh chip to FHT processor engine **120** through MUX **124**.

FIG. 11 shows a detailed block diagram of despreader **178**. Even Walsh chip latch **220** and odd Walsh chip latch **222** latch the data from even Walsh chip buffer **168** and odd Walsh chip buffer **170** respectively. MUX bank **224** extracts the Walsh chip of samples to be used from the two Walsh chips worth of samples presented by even and odd Walsh chip latches **220** and **222**. MUX select logic **226** defines the boundary of the selected Walsh chip based on the offset of the rake element being processed. A Walsh chip is output to OQPSK despreader XOR bank **228**.

The PN sequence values from PN sequence buffer **176** are latched by PN sequence latch **234**. Barrel shifter **232** rotates the output of PN sequence latch **234** based on the offset of the rake element being processed and passes the PN sequence to OQPSK despreader XOR bank **228** which conditionally inverts the antenna samples based on the PN sequence. The XOR'd values are then summed through adder tree **230** which performs the sum operation in the OQPSK despread, and then sums four despread chip outputs together to form a Walsh chip for input to FHT processor engine **120**.

Referring again to FIG. 5, FHT processor engine **120** takes sixty-four received Walsh chips from despreader **178** through MUX **124**, and using a 6-stage butterfly trellis, correlates these sixty-four input samples with each of the sixty-four Walsh functions in a sixty-four clock cycle time slice. Max detect **160** can be used to find the largest of the correlation energies output from FHT processor engine **120**. The output of MAX detect **160** is passed on to search result processor **162** which is part of integrated search processor **128**.

Search result processor **162** is detailed in FIG. 12. Search result processor **162** also operates in a time sliced manner. The control signals provided to it are pipeline delayed to match the two time slice delay from the start of inputting Walsh chips to FHT processor engine **120** to obtaining the maximum energy output. As explained above, a set of search window parameters may designate that a number of Walsh symbols worth of data be accumulated before the results of the chosen offset are processed. In the parameters used with the example of FIGS. 6, 7, 8, and 9, the number of symbols to accumulate is 2. Search result processor **162** performs the summing function along with other functions.

As search result processor **162** performs the sums over consecutive Walsh symbols, it must store a cumulative sum for each rake element in the search rake. These cumulative sums are stored in Walsh symbol accumulation RAM **240**. The results of each search rake are input to summer **242** from max detect **160** for each rake element. Summer **242** sums the present result with the corresponding intermediate value available from Walsh symbol accumulation RAM **240**. On the final Walsh symbol accumulation for each rake element, the intermediate result is read from Walsh symbol accumulation RAM **240** and summed by summer **242** with the final energy from that rake element to produce a final search result for that rake element offset. The search results are then compared with the best results found in the search up to this point as explained below.

In the above mentioned co-pending U.S. Patent Application Serial No. 08/144,902 entitled "DEMODULATION ELEMENT ASSIGNMENT IN A SYSTEM CAPABLE OF RECEIVING MULTIPLE SIGNALS," the preferred embodiment assigns the demodulation elements based on the best results from a search. In the present preferred embodiment, the eight best results are stored in best result register **250**. (A lesser or greater number of results could be stored in other embodiments.) Intermediate result register **164** stores the peak values and their corresponding rank order. If the current search result energy exceeds at least one of the energy values in intermediate result register **164**, search result processor control logic **254** discards the eighth best result in intermediate result register **164**, and inserts the new result, along with its appropriate rank, the PN offset, and antenna corresponding to the rake element result. All lesser ranked results are "demoted" one ranking. There are a great number of methods well known in the art for providing such a sorting function. Any one of them could be used within the scope of this invention.

Search result processor **162** has a local peak filter basically comprised of comparator **244** and previous energy latch **246**. The local peak filter, if enabled, prevents intermediate result register **164** from being updated even though a search result energy would otherwise qualify for inclusion, unless the search result represents a local multipath peak. In this way, the local peak filter prevents strong, broad "smeared" multipath from filling multiple entries in intermediate result register **164**, leaving no room for weaker but distinct multipath that may make better candidates for demodulation.

The implementation of the local peak filter is straightforward. The energy value of the previous rake element summation is stored in previous energy latch **246**. The present rake element summation is compared to the stored value by comparator **244**. The output of comparator **244** indicates which of its two inputs is larger and is latched in search result processor control logic **254**. If the previous sample represented a local maxima, search result processor control logic **254** compares the previous energy result with the data stored in intermediate result register **164** as described above. If the local peak filter is disabled by channel element microprocessor **136** then the comparison with intermediate result register **164** is always enabled. If either the leading or the last rake element at the search window boundary has a slope, then the slope latch is set so the boundary edge value can be considered as a peak as well.

The simple implementation of this local peak filter is aided by the progression of the reads toward earlier symbols within a search rake. As illustrated in FIGS. 6, 7, 8, and 9, within a search rake each rake element progress toward signals arriving earlier in time. This progression means that within a search window, the last rake element of a search rake and the first rake element of the subsequent search rake are contiguous in offset. Therefore, the local peak filter operation does not have to change and the output of comparator **244** is valid across search rake boundaries.

At the end of processing a search window, the values stored in intermediate result register **164** are transferred to best result register **250** readable by channel element microprocessor **136**. Search result processor **162** has thus taken much of the workload from channel element microprocessor **136**, which in the system of FIG. 2 needed to handle each rake element result independently.

The preceding sections have focused on the processing data path of integrated search processor **128** and have detailed how raw antenna samples **118** are translated into a summary multipath report at the output of

best result register **250**. The following sections detail how the each of the elements in the search processing data path are controlled.

Search control block **166** of FIG. 5 is detailed in FIG. 13. As mentioned previously, channel element microprocessor **136** specifies a search parameter set including the group of antennas to search over as stored in antenna select buffer **348**, the starting offset as stored in search offset buffer **308**, the number of rake elements per search rake as stored in rake width buffer **312**, the width of the search window as stored in search width buffer **314**, the number of Walsh symbols to accumulate as stored in Walsh symbol accumulation buffer **316**, and a control word as stored in control word buffer **346**.

The starting offset stored in search offset buffer **308** is specified with eighth chip resolution. The starting offset controls which samples are removed by decimation by latch **218** of FIG. 10 in searcher front end **174**. Due to the two Walsh symbol wide antenna sample buffer **172** in this embodiment, the largest value of the starting offset is half of a PN chip less than two full Walsh symbols.

Up until this point, the generic configuration to perform a search has been disclosed. In reality there are several classes of predefined searches. When a remote unit initially attempts to access the system, it sends a beacon signal called a preamble using the Walsh zero symbol. Walsh zero symbol is the Walsh symbol which contains all logical zeros instead of half ones and zeroes as described above. When a preamble search is performed, the searcher looks for any remote unit sending a Walsh zero symbol beacon signal on an access channel. The search result for a preamble search is the energy for the Walsh zero symbol. When an acquisition mode access channel search is performed, max detect **160** outputs the energy for Walsh zero symbol regardless of the maximum output energy detected. The control word stored in control word buffer **346** includes a preamble bit which indicates when a preamble search is being performed.

As discussed above, the power control mechanism of the preferred embodiment measures the signal level received from each remote unit and creates a power control indication to command the remote unit to raise or lower the remote unit's transmit power. The power control mechanism operates over a set of Walsh symbols called a power control group during traffic channel operation. (Traffic channel operation follows access channel operation and implies operation during an active call.) All the Walsh symbols within a single power control group are transmitted using the same power control indication command at the remote unit.

Also as described above, in the preferred embodiment of the present invention, the signal transmitted by the remote unit is of a variable rate during traffic channel operation. The rate used by the remote unit to transmit the data is unknown at the base station during the searching process. As the consecutive symbols are accumulated, it is imperative that the transmitter is not gated off during the accumulation. Consecutive Walsh symbols in a power control group are gated as a group meaning that the 6 Walsh symbols comprising a power control group in the preferred embodiment are all gated on or all gated off.

Thus when the search parameter specifies that a plurality of Walsh symbols be accumulated during traffic channel operation, the searching process must align each search rake to begin and end within a single power control group. The control word stored in control word buffer **346** includes a power control group alignment bit. With the power control group alignment bit set to one indicating a traffic channel search, the searching process synchronizes to the next power control group boundary instead of just the next offset Walsh symbol boundary.

The control word stored in control word buffer **346** also includes the peak detection filter enable bit as discussed earlier in conjunction with FIG. 8.

The searcher operates either in continuous or single step mode, according to the setting of the continuous/single step bit of the control word. In single step mode, after a search is performed, integrated search processor **128** returns to an idle state to await further instructions. In continuous mode, integrated search processor **128** is always searching, and by the time channel element microprocessor **136** is signaled that the results are available, integrated search processor **128** has started the next search.

Search control block **166** produces the timing signals used to control the searching process performed by integrated search processor **128.** Search control block **166** sends the zero offset timing reference to short code I and Q PN generators **202, 206** and long code User PN generator **204**, and the enable signal to decimator latch **218** and the select signal to MUX **216** in searcher front end **174**. It provides the read and write addresses for PN sequence buffer **176** and even and odd Walsh chip buffers **168** and **170**. It outputs the current offset to control the operation of despreader **178**. It provides the intra-time slice timing reference for FHT processor engine **120**, and determines whether the searching process or the demodulation process uses FHT processor engine **120** by controlling FHT input MUX **124**. It provides several pipeline delayed versions of certain internal timing strobes to search result processor control logic **254** of FIG. 12 to allow it to sum search results

across a rake of offsets for a number of Walsh symbol accumulations. Search control block **166** provides best result register **250** with the pipelined offset and antenna information corresponding to accumulated energy values stored.

In FIG. 13, system time count **342** is slaved to the zero offset time reference. In the preferred embodiment as previously detailed, the system clock runs at eight times the PN chip rate. There are **256** PN chips in a Walsh symbol, and 6 Walsh symbols in a power control group for a total of $6 \times 256 \times 8 = 12{,}288$ system clocks per power control group. Therefore in the preferred embodiment, system time count **342** is comprised of a fourteen bit counter that counts the 12,288 system clocks. The input reference for short code I and Q PN generators **202, 206** and long code User PN generator **204** of FIG. 10 in searcher front end **174** is taken from system time count **342**. (Long code User PN generator **204** output is also based on a longer system wide reference which does not repeat for approximately 50 days. The longer system wide reference is not controlled by the searching process and acts as a preset value. The continuing operation based on the preset value is controlled by system time count **342**.) The addresses for PN sequence buffer **176** and even and odd Walsh chip buffers **168** and **170** are taken from system time count **342**. System time count **342** is latched by latch **328** at the beginning of each time slice. The output of latch **328** is selected via address MUX's **330, 332,** and **334** which provide the write addresses corresponding to the current time slice when these buffers are written at some latter time within the time slice.

Offset accumulator **310** keeps track of the offset of the rake element currently being processed. The starting offset as stored in search offset buffer **308** is loaded into offset accumulator **310** at the beginning of each search window. Offset accumulator **310** is decremented with each rake element. At the end of each search rake that is to be repeated for further accumulations, the number of rake elements per search rake as stored in rake width buffer **312** is added back to the offset accumulator to reference it back to the first offset in the search rake. In this way, the searching process again sweeps across the same search rake for another Walsh symbol accumulation. If the searching process has swept across the current search rake on its final Walsh symbol accumulation then offset accumulator **310** is decremented by one by selection of the "-1" input of repeat rake MUX **304** which produces the offset of the first rake element in the next search rake.

The output of offset accumulator **310** always represents the offset of the current rake element being processed and thus is used to control data input to despreader **178**. The output of offset accumulator **310** is added by adders **336** and **338** to the intra-time slice timing output of system time count **342** to

generate the address sequence within a time slice corresponding to a rake element. The output of adders **336** and **338** is selected via address MUX's **330** and **332** to provide antenna sample buffer **172** read addresses.

The output of offset accumulator **310** is also compared by comparator **326** with the output of system time count **342** to form the offset Walsh symbol strobe which indicates that antenna sample buffer **172** has sufficient valid data for the searching process to begin.

Search rake count **320** keeps track of the number of rake elements remaining to be processed in the current search rake. Search rake count **320** is loaded with the width of the search window as stored in search width buffer **314** at the beginning of a search window. Search rake count **320** is incremented after the processing of the final Walsh symbol accumulation of each search rake is complete. When it reaches its terminal count, all offsets in the search window have been processed. To provide a indication that the end of the current search window is imminent, the output of search rake count **320** is summed by summer **324** with the output of rake width buffer **312**. The end of the search window indication marks the time at which antenna sample buffer **172** may begin to be filled with data samples from an alternative antenna in preparation for the next search window without disrupting the contents needed for the current search window.

When channel element microprocessor **136** specifies a search window, it can specify that the search window be performed for a plurality of antennas. In such a case, the identical search window parameters are repeated using samples from a series of antennas. Such a group of search windows is called an antenna search set. If an antenna search set is specified by channel element microprocessor **136**, the antenna set is programmed by the value stored in antenna select buffer **348**. After the completion of an antenna search set, channel element microprocessor **136** is alerted.

Rake element count **318** contains the number of rake elements left to process in the current search rake. Rake element count **318** is incremented once for each rake element processed and is loaded with the output of rake width buffer **312** when the searching process is in the idle state or upon completion of a search rake.

Walsh symbol accumulation count **322** counts the number of Walsh symbols left to accumulate for the current search rake. The counter is loaded with the number of Walsh symbols to accumulate as stored in Walsh symbol accumulation buffer **316** when the searching process is in the idle state or after completing a search rake sweep on the final Walsh symbol accumulation.

Otherwise the counter is incremented with the completion of each search rake.

Input valid count **302** is loaded whenever the input antenna or decimator alignment changes. It is loaded with the minimum number of samples the searching process needs to process a search rake based on the output of rake width buffer **312** (i.e. one Walsh symbol plus one rake width worth of samples). Each time an antenna sample is written to antenna sample buffer **172**, input valid count **302** is incremented. When it reaches its terminal count, it sends an enable signal that allows the searching process to begin. Input valid count **302** also provides the mechanism for holding off search processing when the offsets of successive search windows do not allow continuous processing of data.

The searching process operates in either an idle state, a sync state, or an active state. Searcher sequencing control **350** maintains the current state. Integrated search processor **128** initializes to the idle state when a reset is applied to channel element modem **110**. During the idle state, all counters and accumulators in search control block **166** load their associated search parameters as presented above. Once channel element microprocessor **136** commands the searching process to begin a continuous or a single step search via the control word, integrated search processor **128** moves to the sync state.

In the sync state, the searching process is always waiting for an offset Walsh symbol boundary. If the data in antenna sample buffer **172** isn't valid yet, or if the power control group alignment bit is set and the Walsh symbol is not a power control group boundary, then integrated search processor **128** remains in the sync state until the proper conditions are met on a subsequent offset Walsh symbol boundary. With a properly enabled offset Walsh symbol, the searching process can move to the active state.

Integrated search processor **128** stays in the active state until it has processed a search rake, at which time it normally returns to the sync state. If integrated search processor **128** is in single step mode, it can go from the active state to the idle state after completing the last rake element for the final Walsh symbol accumulation for the last search rake in the search window. Integrated search processor **128** then waits for channel element microprocessor **136** to initiate another search. If instead, integrated search processor **128** is in continuous mode then at this point it loads the new search parameter set and returns to the sync state to await the offset Walsh symbol at the initial offset to be processed in the new search. The active state is the only state in which the antenna data samples are processed. In the idle or sync states the searching process simply keeps track of time with system time

count **342** and continues to write into the PN sequence buffer **176** and antenna sample buffer **172** so that when the searching process does move to the active state these buffers are ready to be used.

FIG. 14 is an exemplary timing view of the first Walsh symbol accumulation of the second search rake in a search window such as search rake **196** shown in FIG. 9. The third Walsh symbol as referenced to the zero offset reference system time clock is shown divided into thirty-two time slices. Searcher state **372** changes from sync to active when the offset Walsh symbol boundary indication corresponding to Walsh symbol 3 indicates that antenna sample buffer **172** is ready with valid samples to process at that offset. During the next available time slice, the first rake element of the search rake is processed. The searching process continues to use each time slice to process a rack element as indicated by an "S" in time slices **374** unless demod front end **122** uses FHT processor engine **120** as indicated by an "D" in time slices **374**. The searching process finishes processing every rake element in the rake and returns to the sync state before the next offset Walsh symbol boundary corresponding to Walsh symbol 4. Also shown is search rake count state **362** being incremented during the active state until it reaches the terminal state, indicating the complete search rake has been processed. Offset count state **364** is shown being incremented between each time slice corresponding to a rake element, so that it may be used to derive the sample buffer offset read address during the time slice. Offset count state **364** is pipelined delayed as offset count for best result register **366**. The offset count **368** is incremented on the final Walsh symbol accumulation **370** pass.

Thus, a single integrated searcher processor configuration, by buffering antenna samples and utilizing a time sliced transform processor, can independently sequence through a search as configured by a search parameter set, analyze the results and present a summary report of the best paths to use for demodulation element reassignment. This reduces the searching related workload of the controlling microprocessor so that a less expensive microprocessor can be used, and also reduces the direct IC costs by allowing a complete channel element modem on a single IC.

The general principles described herein can be used in systems using alternative transmission schemes. The discussion above was based on the reception of a reverse link signal where no pilot signal is available. On the forward link of the preferred embodiment, the base station transmits a pilot signal. The pilot signal is a signal having known data thus the FHT process used to determine which data was transmitted is not necessary. In order to embody the present invention, a integrated search processor for receiving a

signal comprising a pilot signal would not contain the FHT processor or maximum detection function. For example FHT processor engine **120** and max detect **160** blocks of FIG. 5 could be replaced with simple accumulator **125** as shown in FIG. 15. The searching operation when a pilot signal is available is analogous to an acquisition mode access channel search operation as described above.

The searching architecture described above can be used to perform searches in a variety of manners. The most efficient search is a linear search. A linear search is performed by linearly searching potential time offsets in order regardless of the probability that the remote unit is transmitting. When searching for a remote unit signal, the base station must know the expected coverage area range. For example a typical base station covers a range of approximately 50 kilometers implying a round trip delay of 350 microseconds or approximately 430 PN chips in the preferred embodiment. Also, in the multipath environment where signals take nondirect paths, the remote unit signal may be delayed as much as twice the direct path propagation implying that searching must be done over a set of nearly 1000 different PN offsets. Once a remote unit's signal has been detected and is being demodulated, the approximate distance of the remote unit is known and the possible PN offsets which need to be searched to ensure that the majority valid multipath signals are detected are greatly reduced.

Within a given search over a power control group, there are three reasons why a signal may not be detected at a given PN offset. First, no signal may be arriving at the given PN offset. A remote unit may provide several multipath signals but the number of multipath signals created is only a very small portion of all the offsets searched. Thus the majority of searched offsets do not produce energy results that exceed the detection threshold precisely because no remote unit signal is present at that offset.

Secondly, the signal may be arriving at the given PN offset but faded during a large portion of the search integration time. As explained above, the multipath characteristic of a radio channel can result in signal fading. Fading is the result of the phasing characteristics of the multipath channel. A fade occurs when multipath vectors are added destructively, yielding a received signal that is smaller than either individual vector. Thus if a signal which is long term valid happens to be in a deep fade at the time the search is made, no signal is available for detection by the searching process.

Thirdly, the signal would have arrived at the given PN offset but for the fact that the transmitter of the remote unit is gated off during the period of time in question. As explained above, in the preferred embodiment the

remote unit produces a bursty signal. The remote unit comprises a variable rate vocoder which produces variable rate frames of data. The data burst randomizer determines during which time periods the remote unit transmits and during which time periods it does not transmit given the data rate of the signal to be transmitted, a remote unit specific identifying number, and the time of day. When operating at less than full rate, the data burst randomizer within the remote unit pseudorandomly distributes the active time periods within the transmission burst. A corresponding data burst randomizer is also included in the base station such that the base station can recreate the pseudorandom distribution based on the time of day and the remote unit specific identifying number but the rate information is not available during the searching process. As noted above, the eighth rate time periods determine a so called worthy group of time periods. In this way, regardless of the data rate of the transmitted signal, each time period corresponding to the worthy group is sure to correspond to a time when the corresponding remote unit was transmitting a signal. During all other time periods, the remote unit may or may not be transmitting depending on the corresponding encoding rate.

When a linear search is specified, in order to obtain valid power measurements, the searching process confines the search integration time (i.e. the number of Walsh accumulations at a single search offset) to begin and end within a single power control group as explained in greater detail above. A search that integrates only within a single power control group is said to be synchronized with the power control group boundaries. If the searching process at a given offset were accumulated without regard to power control group boundaries and the remote unit were transmitting at less than full rate, valid search results corresponding to a power control group where the remote unit's signal is gated on may be summed with noise accumulated during a subsequent power control group that remote unit's signal is gated off. The summation of the search results corresponding to the power control group where the remote unit's signal is gated off corrupt the otherwise valuable results accumulated during the power control that the remote unit's signal is gated on.

One method of searching would be to search only those power control groups corresponding to worthy groups. Even if such worthy group only searching is performed, the searching process and demodulation element assignment process must still be capable of handling the situation in which the energy accumulated does not exceed the detection threshold but in reality a signal is present at the offset due to the unpredictable fading characteristics

of the channel. Therefore a more efficient scheme is to accumulate energy in all power control groups whether or not they correspond to worthy groups. If energy is detected in a search which does not correspond to a worthy group, an additional valid data point is generated over and above what would be generated based on a worthy group only search.

As noted above, a preamble search and a search performed during traffic channel operation are different. When a remote unit initially attempts to access the system, it sends a beacon signal called a preamble using the Walsh zero symbol. Walsh zero symbol is the Walsh symbol which contains all logical zeros instead of half ones and zeroes as described above. When a preamble search is performed, the searcher looks for any remote unit sending a Walsh zero symbol beacon signal on an access channel. In the preferred embodiment, the transmission of the preamble is always full rate and is never gated off. Therefore during a preamble search there is no need for synchronization with the power control group boundaries.

There are many configurations for spread spectrum multiple access communication systems not specifically described herein but with which the present invention is applicable. For example, other encoding and decoding means could be used instead of the Walsh encoding and FHT decoding. The previous description of the preferred embodiments is provided to enable any person skilled in the art to make or use the present invention. The various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without the use of the inventive faculty. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WE CLAIM:

CLAIMS

1. A method of receiving a signal comprised of a group of spread spectrum call signals sharing a common frequency band wherein each of said spread spectrum call signals comprises a series of bits encoded in groups of a fixed length into a series of symbols wherein a series of said symbols are grouped together in a power control group wherein each symbol in a common power control group is transmitted at a common power level and wherein said power control groups are transmitted in bursts, and isolating one of said call signals from among said group to determine a call signal strength at a path delay time offset from a zero offset reference time, said method comprising the steps of:

storing PN sequence data bits in a PN sequence buffer;

storing a first received set of call signal samples in a sample buffer having a limited size;

despreading a first fixed length set of said call signal samples from said sample buffer corresponding to a first path delay time with a first set of PN sequence data bits from said PN sequence buffer to produce a first despread output;

storing a second received set of call signal samples in said sample buffer; and

despreading a second fixed length set of call signal samples from said sample buffer corresponding to a second path delay time with said first set of PN sequence data bits from said PN sequence buffer to produce a second despread output;

wherein said second fixed length set of call signal samples comprises a large number of the same call signal samples as said first fixed length set of call signal samples and wherein the length of said first and second received set of call signal samples is a fraction the fixed length of said first and second fixed length set of call signal samples;

wherein said steps of storing said first and second fixed length set of call signal samples and said steps of despreading said first and second fixed length set of calls signal samples are performed independent of a probability that said one of said call signals is comprises one of said power control groups.

2. A method of receiving a signal comprised of a group of spread spectrum signals sharing a common frequency band and isolating a first signal from among said group of spread spectrum signals to determine a signal strength at a path delay time offset from a zero offset reference time of said first signal wherein said first signal comprises a series of symbols wherein a series of said symbols are grouped together in a symbol set wherein each symbol in a common symbol set is transmitted at a fixed power level wherein successive symbol sets may be transmitted at a variety of signal levels wherein said variety of signal levels includes a zero level wherein transmission of said first signal is gated off, said method comprising the steps of:

searching a first set of call signal samples corresponding to a first symbol set for said first signal at a first offset to produce a first power estimate thereof;

searching a second set of call signal samples corresponding to said first symbol set for said first signal at said first offset to produce a second power estimate thereof;

summing said first and second power estimates to produce a symbol set power level estimate at said first offset;

searching a third set of call signal samples corresponding to a second symbol set for said first signal at a second offset to produce a third power estimate thereof;

searching a fourth set of call signal samples corresponding to said second symbol set for said first signal at said second offset to produce a fourth power estimate thereof; and

summing said third and fourth power estimates to produce a symbol set power level estimate at said second offset;

wherein said first symbol set and said second symbol set correspond to time contiguous symbol sets and wherein said steps of searching are performed continually regardless of said fixed power level.

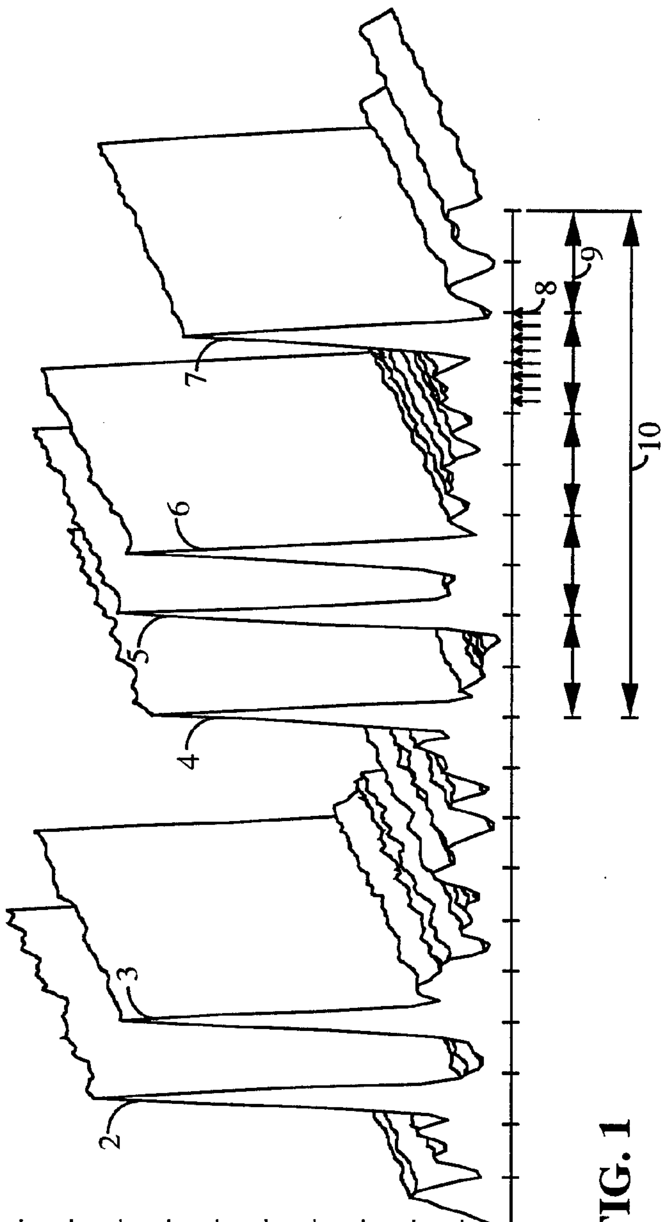

FIG. 1

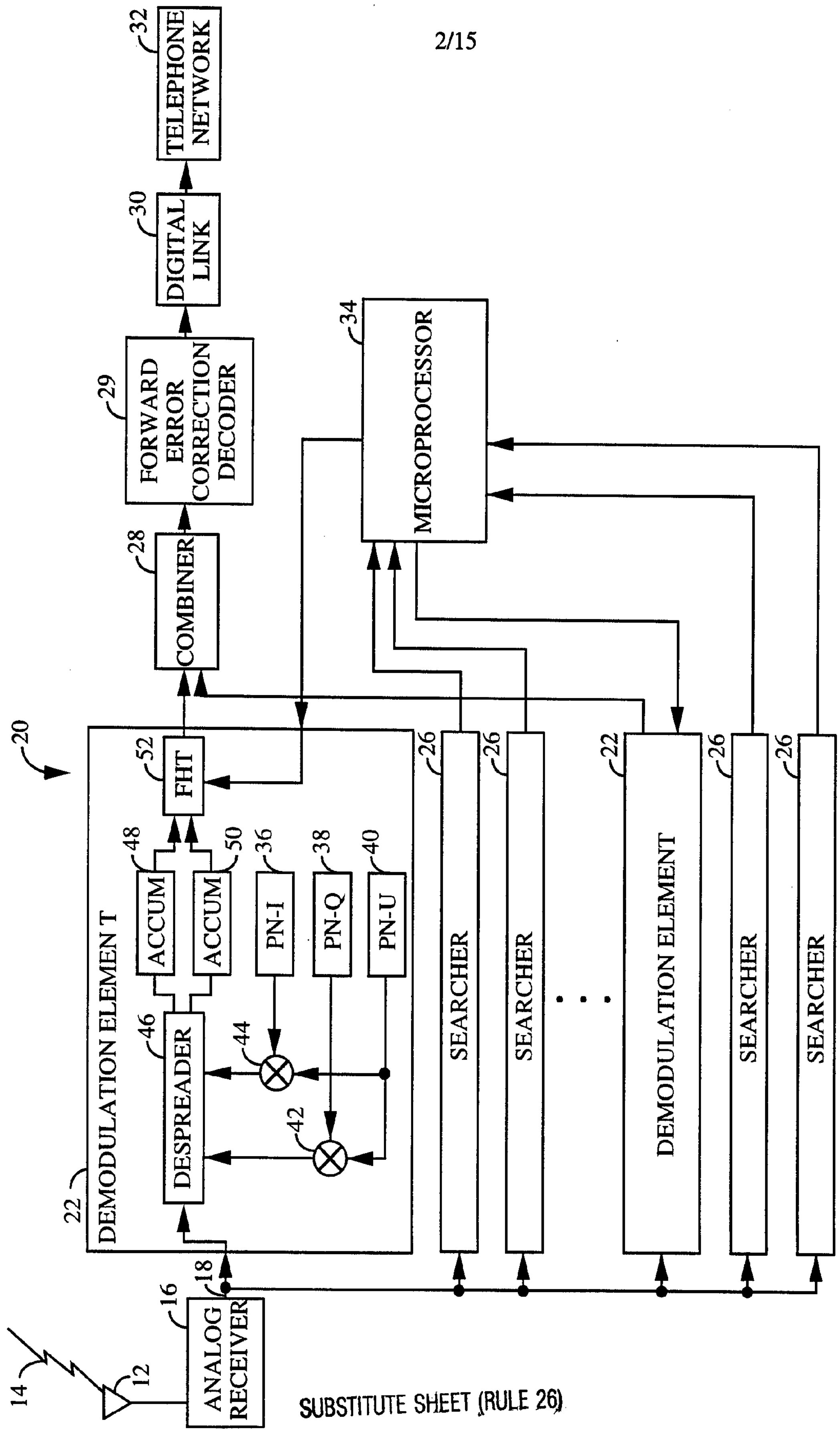
TELEPHONE NETWORK
32
DIGITAL LINK
30
FORWARD ERROR CORRECTION DECODER
29
COMBINER
28
MICROPROCESSOR
34
20
DEMODULATION ELEMEN T
22
FHT
52
ACCUM
48
ACCUM
50
PN-I
36
PN-Q
38
PN-U
40
DESPREADER
46
44
42
SEARCHER
26
SEARCHER
26
DEMODULATION ELEMENT
22
SEARCHER
26
SEARCHER
26
ANALOG RECEIVER
16
18
12
14

FIG. 2

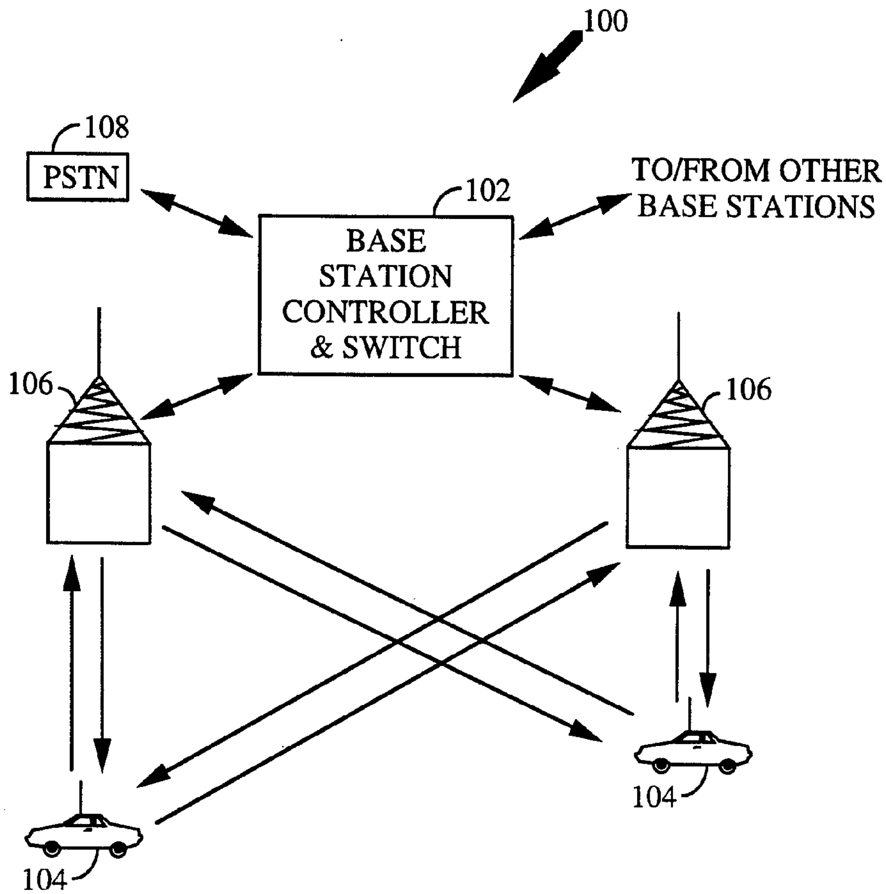
100
108
PSTN
102
BASE
STATION
CONTROLLER
& SWITCH
TO/FROM OTHER
BASE STATIONS
106
106
104
104

FIG. 3

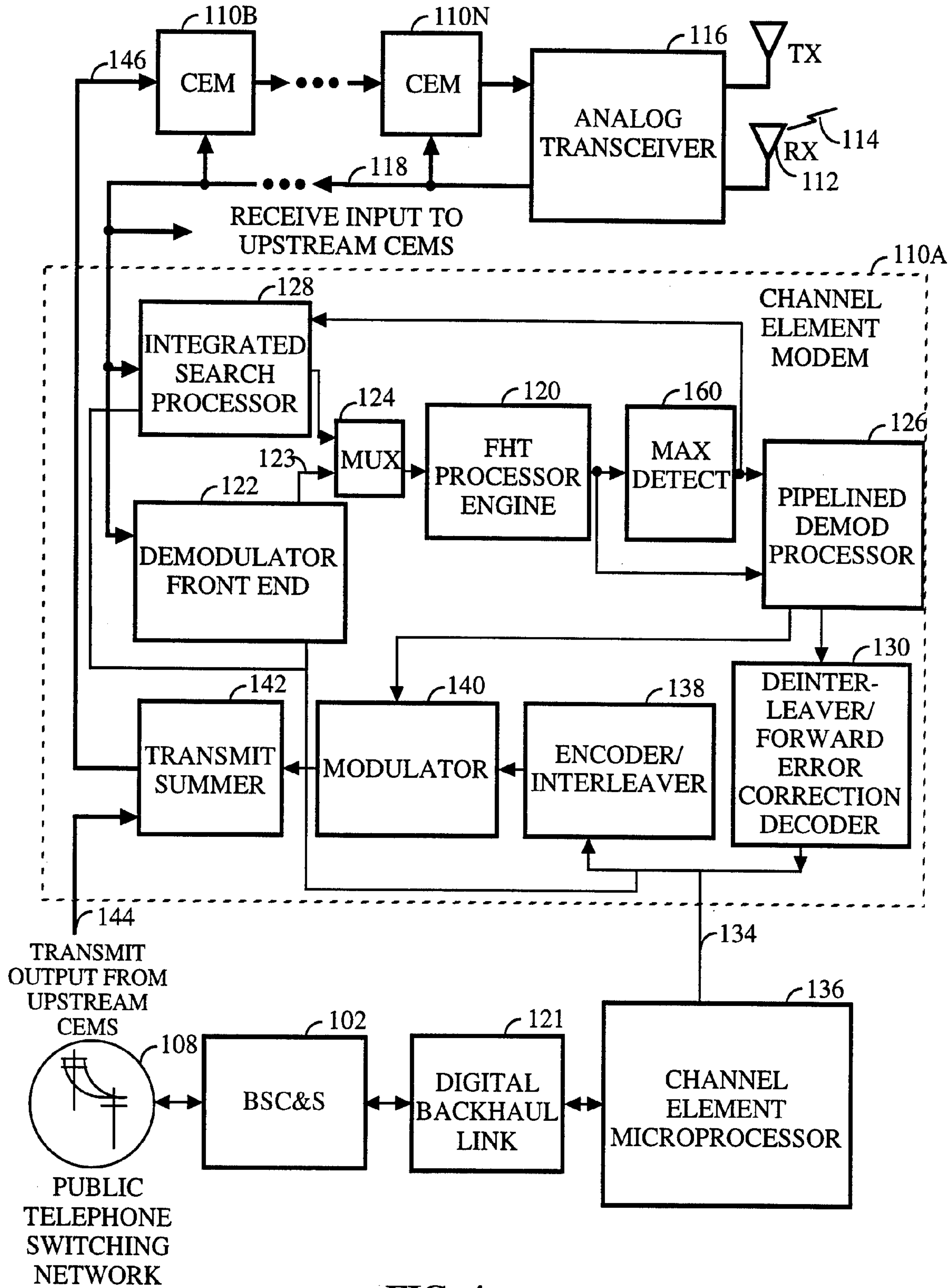
110B
CEM
110N
CEM
116
TX
ANALOG
TRANSCEIVER
114
RX
112
146
118
RECEIVE INPUT TO
UPSTREAM CEMS
110A
CHANNEL
ELEMENT
MODEM
128
INTEGRATED
SEARCH
PROCESSOR
124
MUX
120
FHT
PROCESSOR
ENGINE
160
MAX
DETECT
126
PIPELINED
DEMOD
PROCESSOR
123
122
DEMODULATOR
FRONT END
130
DEINTER-
LEAVER/
FORWARD
ERROR
CORRECTION
DECODER
142
TRANSMIT
SUMMER
140
MODULATOR
138
ENCODER/
INTERLEAVER
144
134
TRANSMIT
OUTPUT FROM
UPSTREAM
CEMS
136
108
102
BSC&S
121
DIGITAL
BACKHAUL
LINK
CHANNEL
ELEMENT
MICROPROCESSOR
PUBLIC
TELEPHONE
SWITCHING
NETWORK

FIG. 4

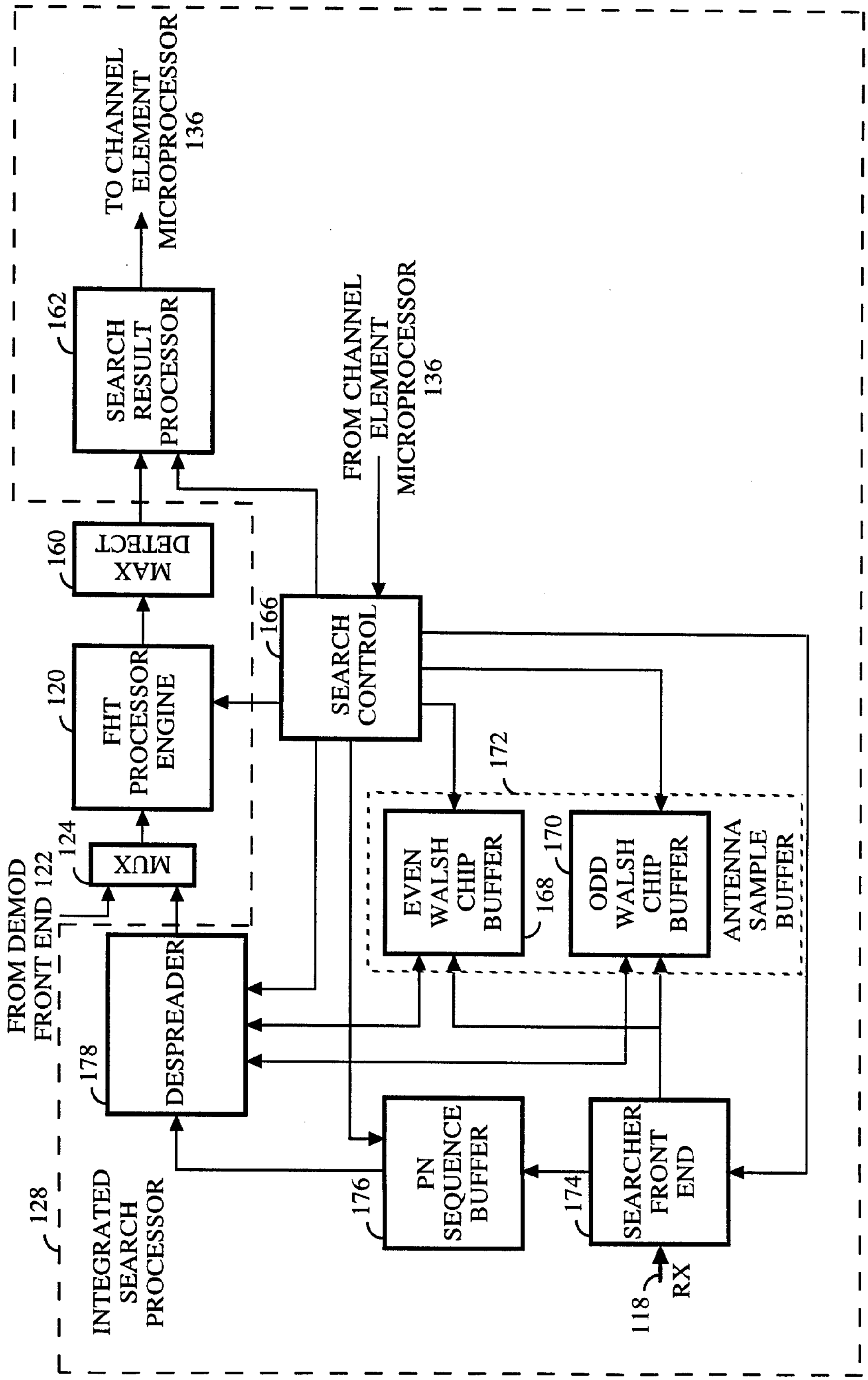
TO CHANNEL ELEMENT MICROPROCESSOR 136
162
SEARCH RESULT PROCESSOR
FROM CHANNEL ELEMENT MICROPROCESSOR 136
160
MAX DETECT
166
SEARCH CONTROL
120
FHT PROCESSOR ENGINE
172
124
MUX
FROM DEMOD FRONT END 122
EVEN WALSH CHIP BUFFER
168
170
ODD WALSH CHIP BUFFER
ANTENNA SAMPLE BUFFER
178
DESPREADER
PN SEQUENCE BUFFER
176
SEARCHER FRONT END
174
128
INTEGRATED SEARCH PROCESSOR
118
RX

FIG. 5

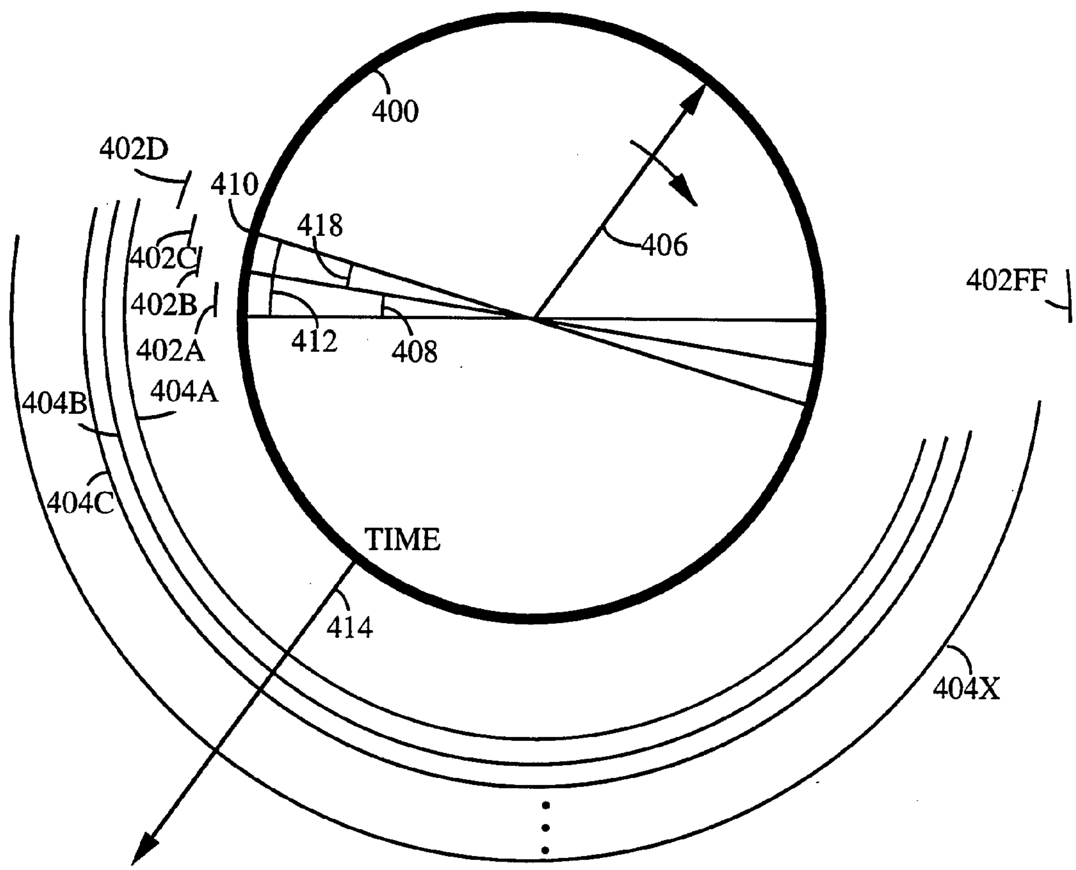
400
402D
410
418
402C
402B
402A
412
408
406
402FF
404B
404A
404C
TIME
414
404X

FIG. 6

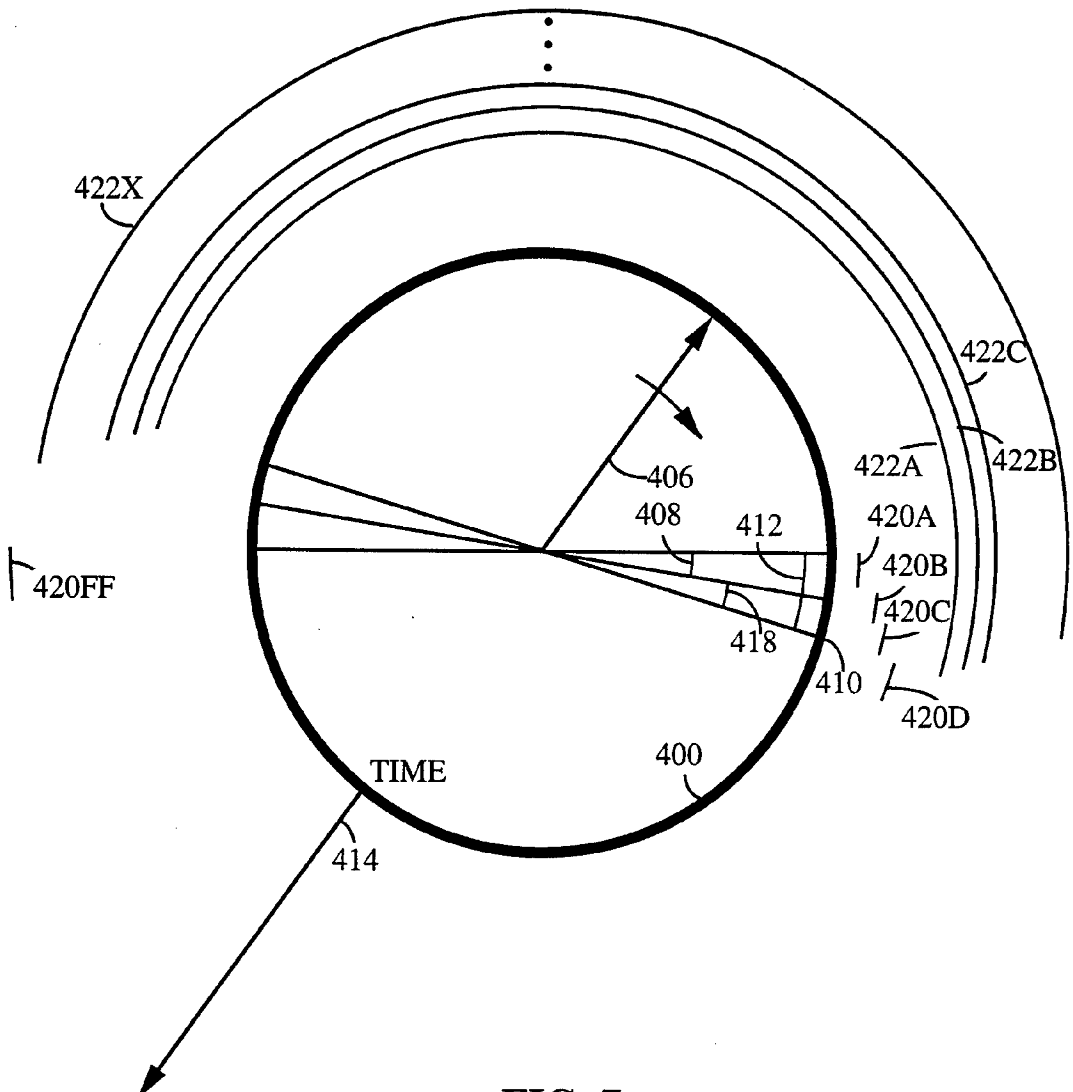
422X
422C
422A
422B
406
408
412
420A
420B
420C
420FF
418
410
420D
400
TIME
414

FIG. 7

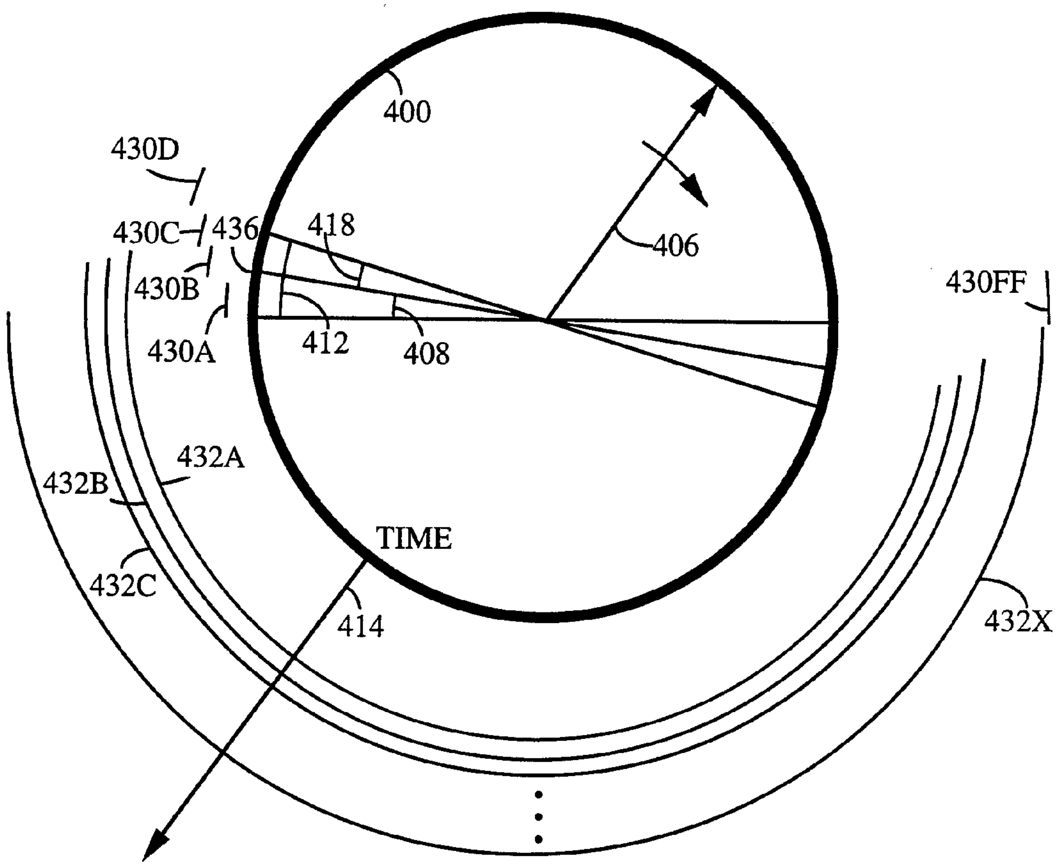
400
430D
430C
436
418
406
430B
430FF
430A
412
408
432A
432B
TIME
432C
414
432X

FIG. 8

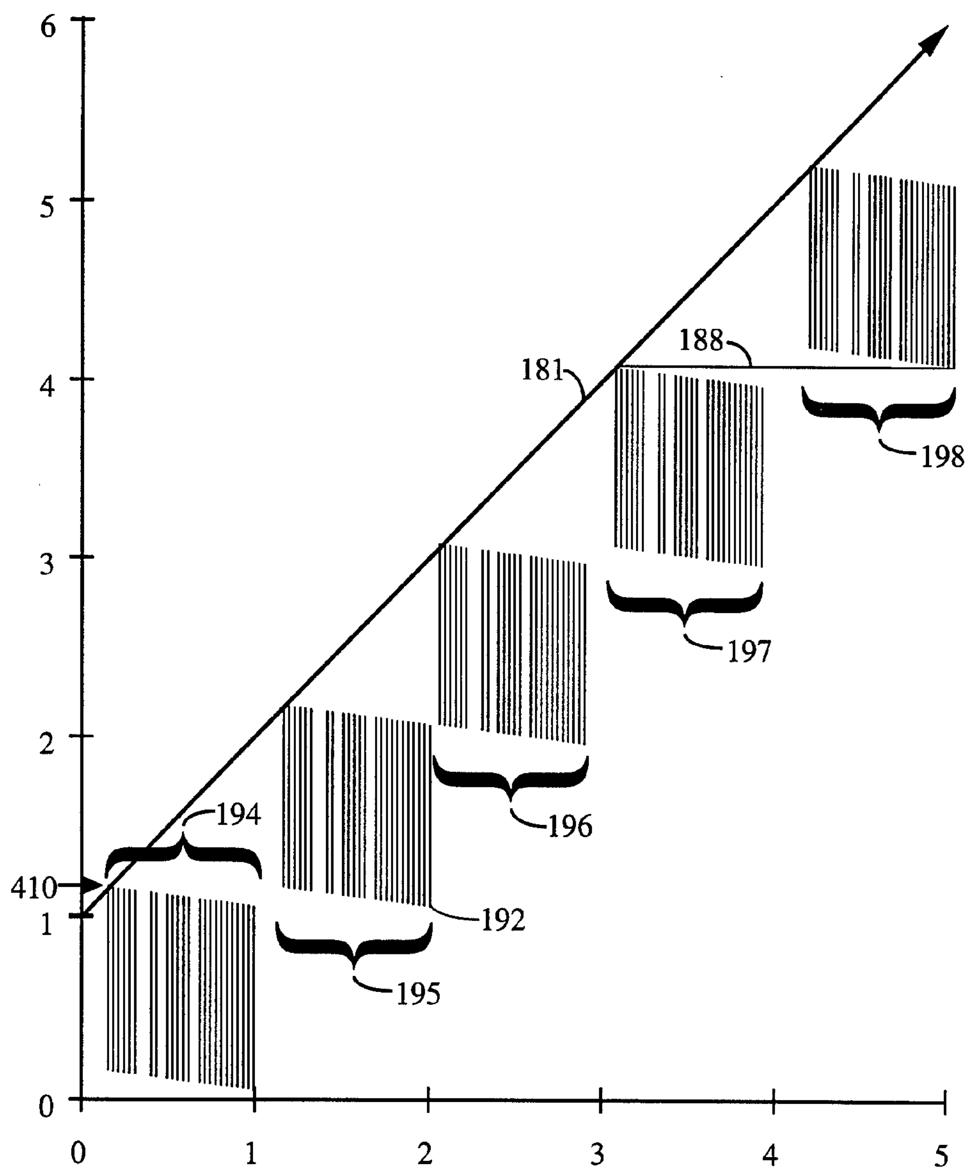

FIG. 9

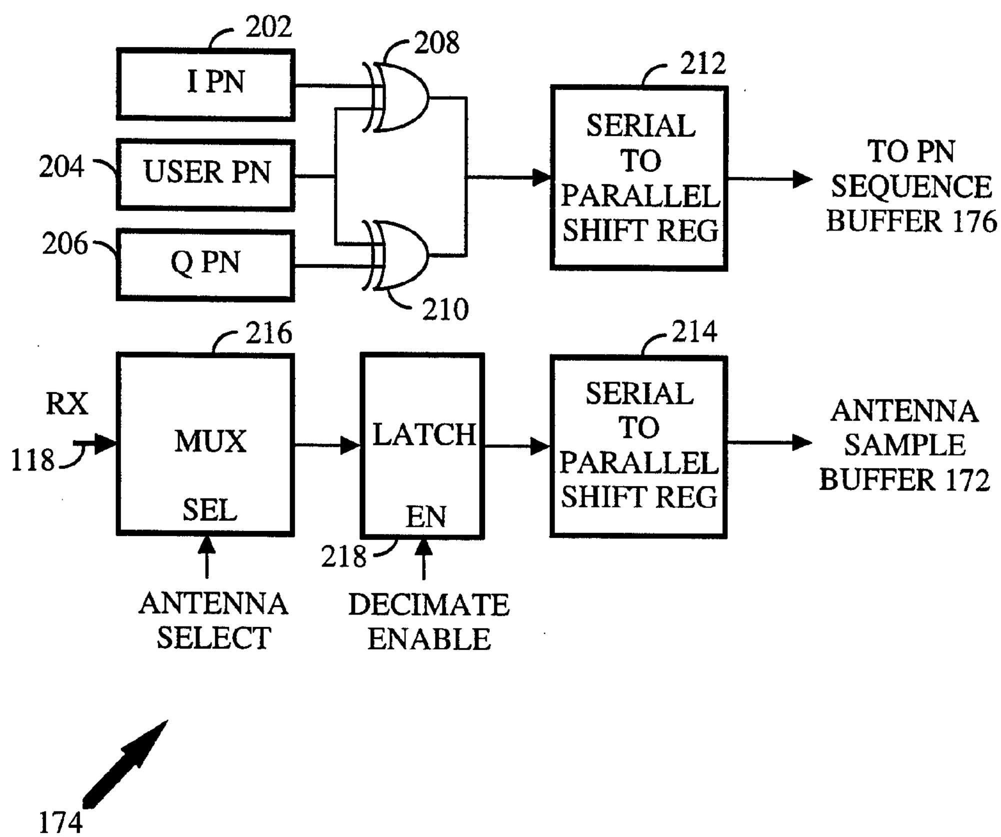
202
I PN
208
212
204
USER PN
SERIAL
TO
PARALLEL
SHIFT REG
TO PN
SEQUENCE
BUFFER 176
206
Q PN
210
216
214
RX
118
MUX
SEL
LATCH
EN
SERIAL
TO
PARALLEL
SHIFT REG
ANTENNA
SAMPLE
BUFFER 172
218
ANTENNA
SELECT
DECIMATE
ENABLE
174

FIG. 10

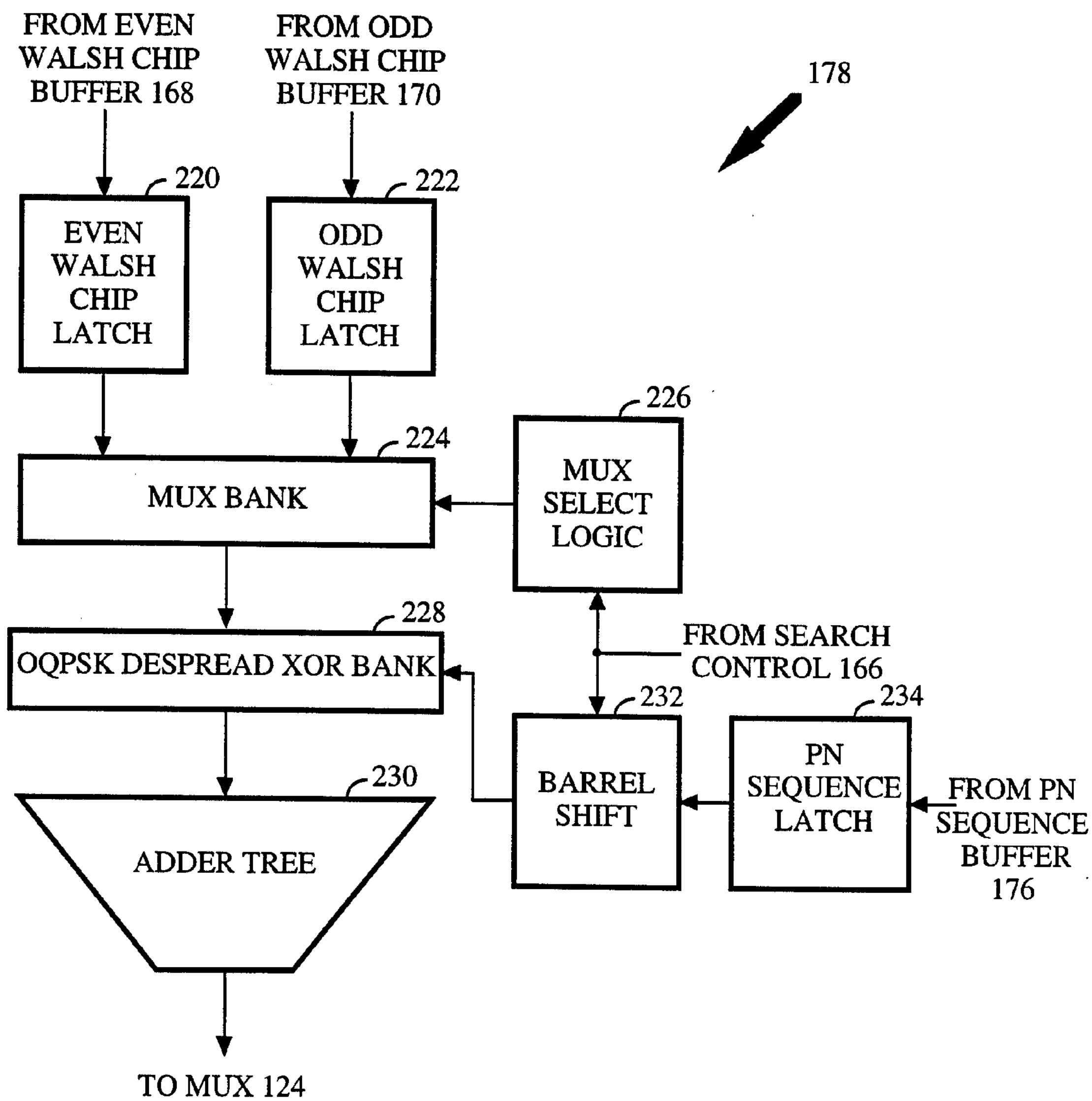
FROM EVEN WALSH CHIP BUFFER 168
FROM ODD WALSH CHIP BUFFER 170
178
220
EVEN WALSH CHIP LATCH
222
ODD WALSH CHIP LATCH
226
MUX SELECT LOGIC
224
MUX BANK
228
OQPSK DESPREAD XOR BANK
FROM SEARCH CONTROL 166
232
234
BARREL SHIFT
PN SEQUENCE LATCH
FROM PN SEQUENCE BUFFER 176
230
ADDER TREE
TO MUX 124

FIG. 11

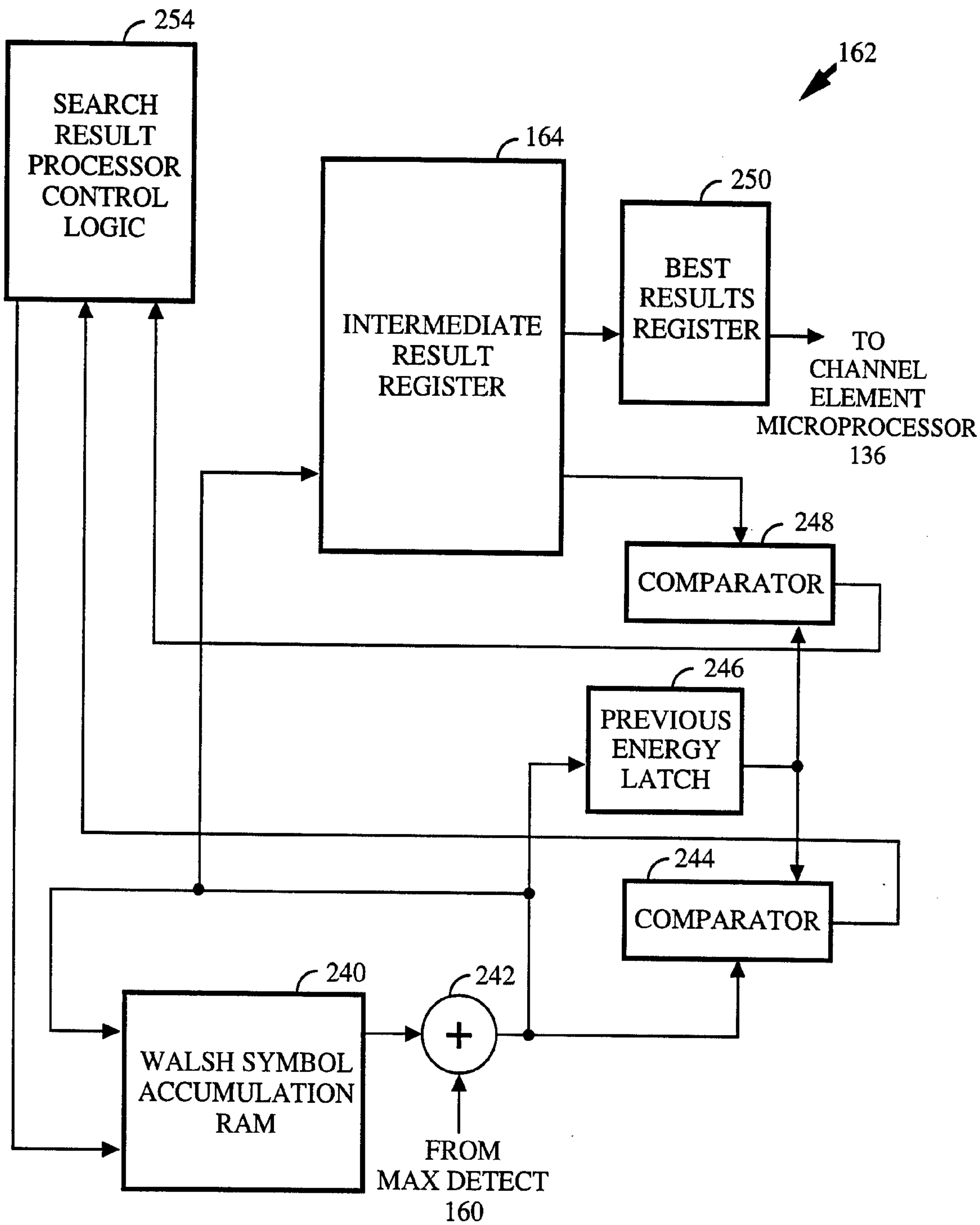
254
SEARCH RESULT PROCESSOR CONTROL LOGIC
162
164
INTERMEDIATE RESULT REGISTER
250
BEST RESULTS REGISTER
TO CHANNEL ELEMENT MICROPROCESSOR 136
248
COMPARATOR
246
PREVIOUS ENERGY LATCH
244
COMPARATOR
240
WALSH SYMBOL ACCUMULATION RAM
242
+
FROM MAX DETECT 160

FIG. 12

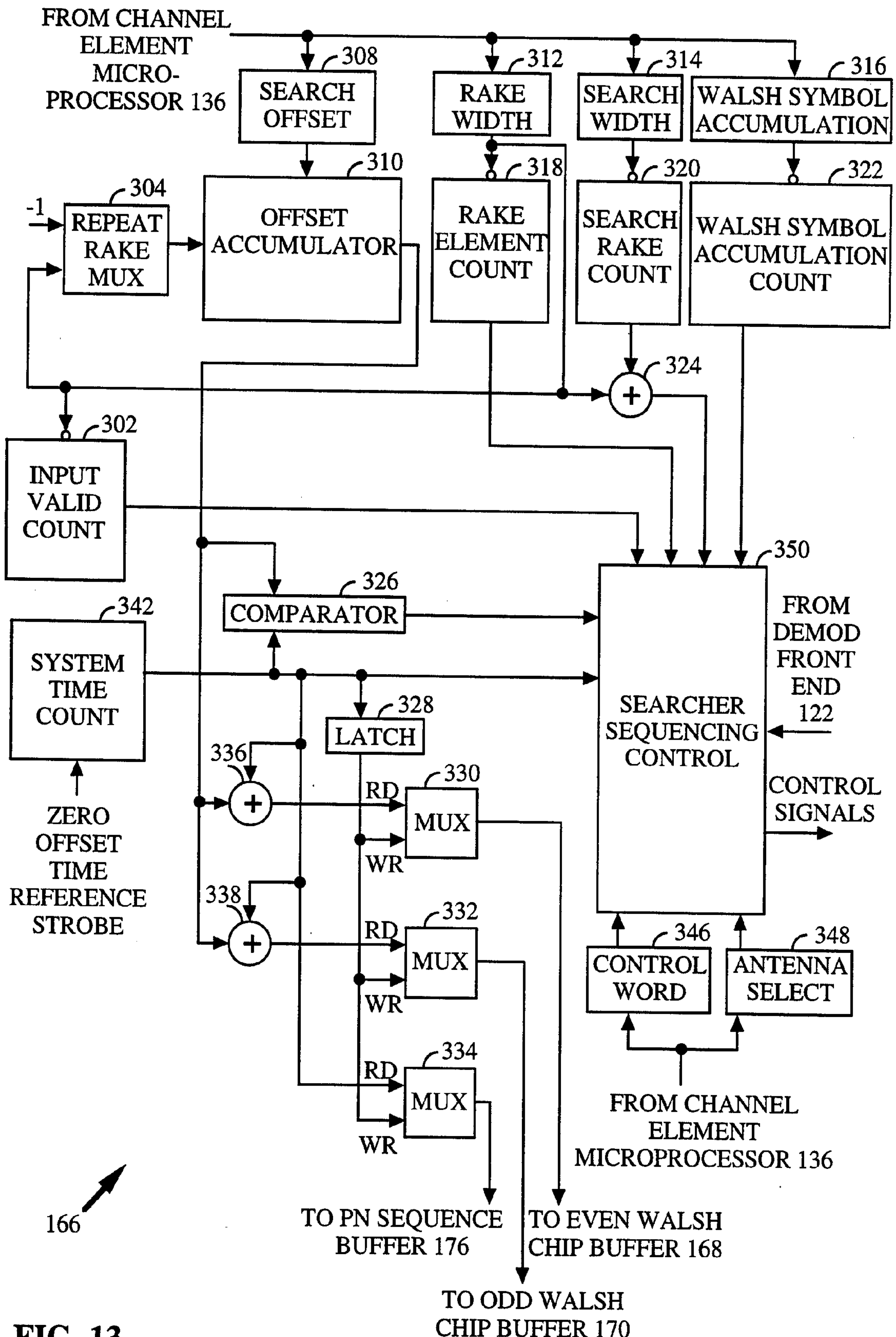
FROM CHANNEL ELEMENT MICRO-PROCESSOR 136
308
SEARCH OFFSET
312
RAKE WIDTH
314
SEARCH WIDTH
316
WALSH SYMBOL ACCUMULATION
310
OFFSET ACCUMULATOR
304
-1
REPEAT RAKE MUX
318
RAKE ELEMENT COUNT
320
SEARCH RAKE COUNT
322
WALSH SYMBOL ACCUMULATION COUNT
324
302
INPUT VALID COUNT
350
326
COMPARATOR
342
SYSTEM TIME COUNT
FROM DEMOD FRONT END 122
328
LATCH
SEARCHER SEQUENCING CONTROL
336
330
RD
MUX
WR
CONTROL SIGNALS
ZERO OFFSET TIME REFERENCE STROBE
338
332
RD
MUX
WR
346
CONTROL WORD
348
ANTENNA SELECT
334
RD
MUX
WR
FROM CHANNEL ELEMENT MICROPROCESSOR 136
166
TO PN SEQUENCE BUFFER 176
TO EVEN WALSH CHIP BUFFER 168
TO ODD WALSH CHIP BUFFER 170

FIG. 13

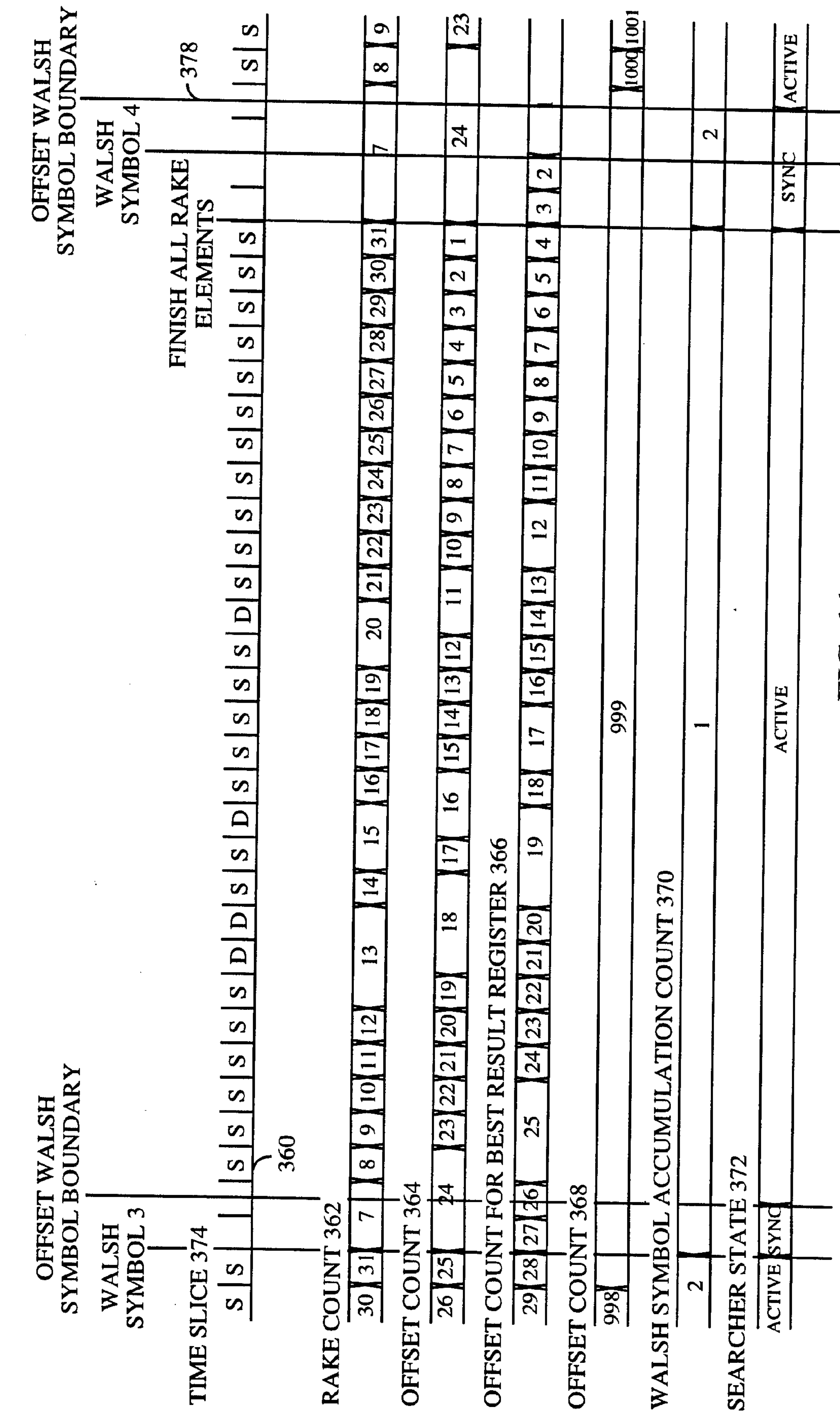
OFFSET WALSH SYMBOL BOUNDARY
WALSH SYMBOL 3
TIME SLICE 374
360
RAKE COUNT 362
OFFSET COUNT 364
OFFSET COUNT FOR BEST RESULT REGISTER 366
OFFSET COUNT 368
WALSH SYMBOL ACCUMULATION COUNT 370
SEARCHER STATE 372
FINISH ALL RAKE ELEMENTS
WALSH SYMBOL 4
OFFSET WALSH SYMBOL BOUNDARY
378
ACTIVE
SYNC
999
998
1000
1001

FIG. 14

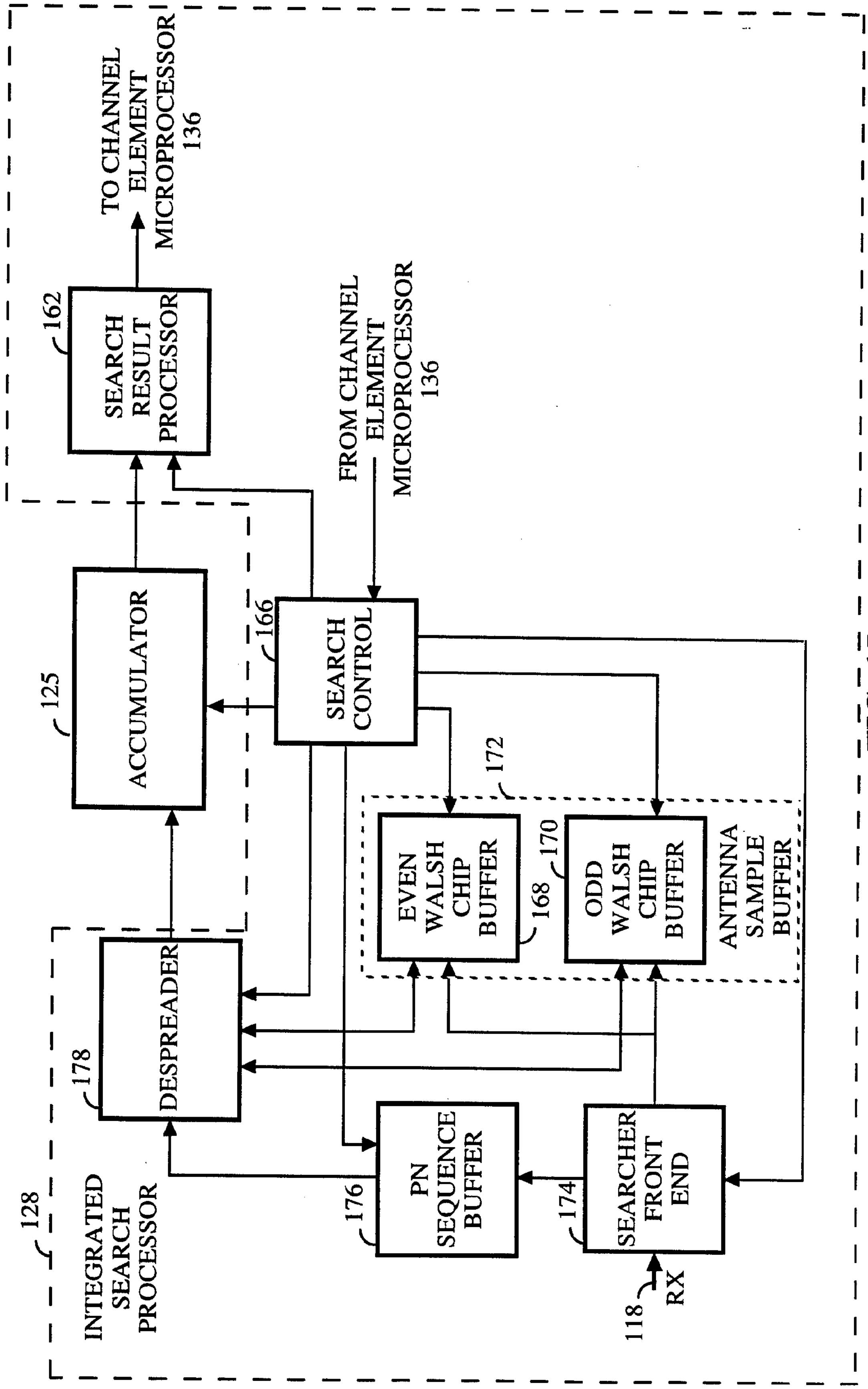
TO CHANNEL ELEMENT MICROPROCESSOR 136
162
SEARCH RESULT PROCESSOR
FROM CHANNEL ELEMENT MICROPROCESSOR 136
125
ACCUMULATOR
166
SEARCH CONTROL
172
EVEN WALSH CHIP BUFFER
168
170
ODD WALSH CHIP BUFFER
ANTENNA SAMPLE BUFFER
178
DESPREADER
128
INTEGRATED SEARCH PROCESSOR
176
PN SEQUENCE BUFFER
174
SEARCHER FRONT END
118
RX

FIG. 15

INTERNATIONAL SEARCH REPORT

In ational Application No
PCT/US 96/07567

A. CLASSIFICATION OF SUBJECT MATTER
IPC 6 H04B7/26 H04B1/707

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6 H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X,P	WO,A,96 10873 (QUALCOMM INC) 11 April 1996 cited in the application see page 26, line 29 - page 29, line 15; claims 1,16,19-21,24; figures 1,5 ---	1,2
A	WO,A,95 01018 (QUALCOMM INC) 5 January 1995 see abstract; claims 1,2 ---	2
A	US,A,5 109 390 (GILHOUSEN KLEIN S ET AL) 28 April 1992 cited in the application see abstract; figure 3 see column 14, line 47 - column 15, line 40 --- -/--	1,2

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

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"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

1

Date of the actual completion of the international search	Date of mailing of the international search report
23 September 1996	- 8. 10. 96

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax (+31-70) 340-3016	Kolbe, W

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 96/07567

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category *	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	GB,A,2 278 983 (ROKE MANOR RESEARCH) 14 December 1994 see abstract; claim 1 -----	1,2

1

INTERNATIONAL SEARCH REPORT

Information on patent family members

In tional Application No
PCT/US 96/07567

Patent document cited in search report	Publication date	Patent family member(s)	Publication date
WO-A-9610873	11-04-96	AU-A- 3945195	26-04-96
		CA-A- 2174243	11-04-96
		EP-A- 0732013	18-09-96
WO-A-9501018	05-01-95	US-A- 5442627	15-08-95
		BR-A- 9406851	05-03-96
		CA-A- 2165801	05-01-95
		CN-A- 1103521	07-06-95
		CN-A- 1125498	26-06-96
		EP-A- 0705510	10-04-96
		FI-A- 956253	22-02-96
		ZA-A- 9404074	06-03-95
US-A-5109390	28-04-92	AU-B- 649987	09-06-94
		AU-A- 6874891	31-05-91
		CA-A- 2072876	08-05-91
		CN-A- 1061311	20-05-92
		EP-A- 0500761	02-09-92
		IL-A- 96220	12-04-94
		JP-T- 4502844	21-05-92
		WO-A- 9107036	16-05-91
GB-A-2278983	14-12-94	CN-A- 1109266	27-09-95
		EP-A- 0653127	17-05-95
		FI-A- 946184	30-12-94
		WO-A- 9428640	08-12-94
		JP-T- 7509831	26-10-95

WO9809381

PUB DATE: 1998-03-05

APPLICANT: UNIV LELAND STANFORD JUNIOR [US]

HAS ATTACHED HERETO CORRESPONDING ENGLISH LANGUAGE EQUIVALENT:

JP2001505723

PUB DATE: 2001-04-24

APPLICANT: Lucent Technologies Inc.Murray Hill N.J

HIGH CAPACITY WIRELESS COMMUNICATION USING SPATIAL SUBCHANNELS

Publication number: WO9809381 (A1)
Publication date: 1998-03-05
Inventor(s): RALEIGH GREGORY G [US]; CIOFFI JOHN M [US] +
Applicant(s): UNIV LELAND STANFORD JUNIOR [US] +
Classification:
- international: ***H04B7/04; H04B7/08; H04B7/24; H04J15/00; H04L1/06; H04L25/02; H04L27/26;*** *H04B1/10; H04B7/06; H04L1/00; H04L1/18; H04W16/28;* (IPC1-7): H04B1/38; H04M1/00
- European: H04B7/04M1; H04B7/06C1F7; H04B7/08C4J2; H04L1/00B7B; H04L1/00B7C; H04L1/00B7K1; H04L1/00B7V; H04L1/06T; H04L25/02C1; H04L25/02C11A; H04L25/02C3; H04L25/03B9; H04L27/26M3; H04L27/26M5
Application number: WO1997US15363 19970829
Priority number(s): US19960025227P 19960829; US19960025228P 19960829

Also published as:

- WO9809395 (A1)
- WO9809385 (A2)
- WO9809385 (A3)
- JP2001505723 (T)
- EP0931388 (A2)

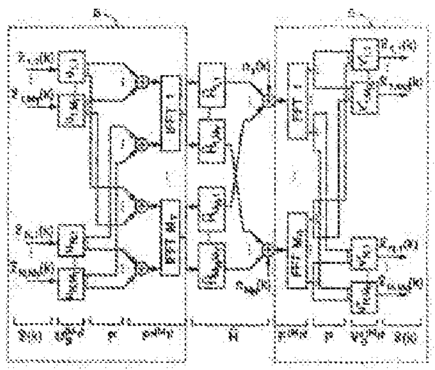
more >>

Cited documents:

- US4710944 (A)
- US5548819 (A)
- US5649287 (A)

Abstract of **WO 9809381 (A1)**

In a system and method of digital wireless communication between a base station (B) and a subscriber unit (S), a spatial channel characterized by a channel matrix H couples an adaptive array of Mt antenna elements (1-Mt) at the base station (B) with an adaptive array Mr antenna elements (1-Mr) at the subscriber unit (S). The method comprises the step of determining from the channel matrix H a number L of independent spatio-temporal subchannels, and encoding a plurality of information signals into a sequence of transmitted signal vectors. The transmitted signal vectors have Mt complex valued components and are selected to transmit distinct signal information in parallel over the independent subchannels, thereby providing increased communication capacity between the base and the subscriber. The sequence of transmitted signal vectors is transmitted from the base station array (1-Mt), and a sequence of received signal vectors is received at the subscriber array (1-Mr) and are decoded to yield the original information signals. Specific spatio-temporal coding techniques are described that increase system performance.



PCT WORLD INTELLECTUAL PROPERTY ORGANIZATION
International Bureau

INTERNATIONAL APPLICATION PUBLISHED UNDER THE PATENT COOPERATION TREATY (PCT)

(51) International Patent Classification 6 : **H04B 1/38, H04M 1/00**	A1	(11) International Publication Number: **WO 98/09381** (43) International Publication Date: 5 March 1998 (05.03.98)

(21) International Application Number: PCT/US97/15363

(22) International Filing Date: 29 August 1997 (29.08.97)

(30) Priority Data:

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60/025,228	29 August 1996 (29.08.96)	US

(71) Applicant: THE BOARD OF TRUSTEES OF THE LELAND STANFORD JUNIOR UNIVERSITY [US/US]; Suite 350, 900 Welch Road, Palo Alto, CA 94304 (US).

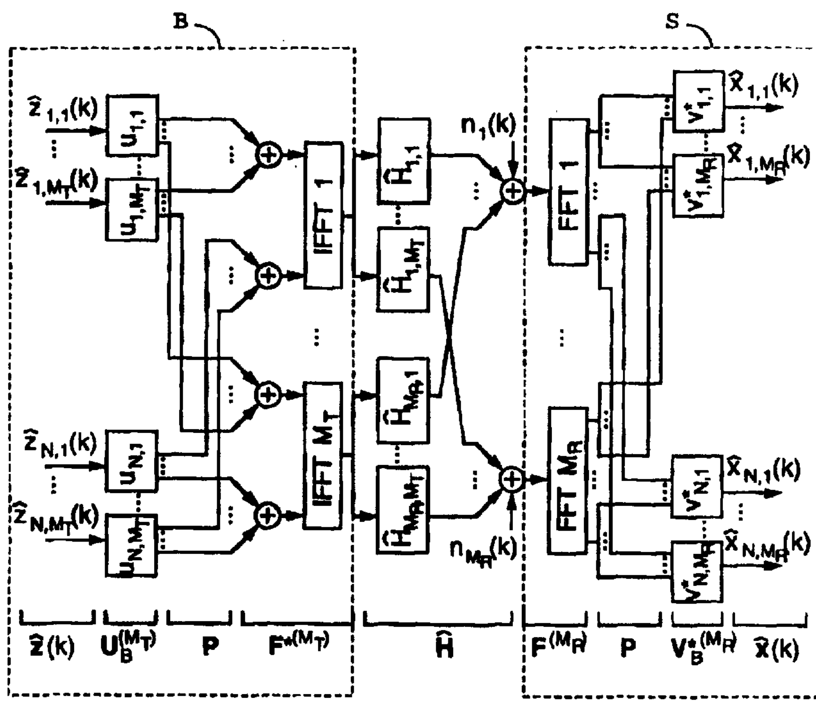
(72) Inventors: RALEIGH, Gregory, G.; 539 San Juan Avenue, El Granada, CA 94018 (US). CIOFFI, John, M.; 14298 Saddle Mountain Drive, Los Altos Hills, CA 94022 (US).

(74) Agent: McFARLANE, Thomas, J.; 426 Lowell Avenue, Palo Alto, CA 94301 (US).

(81) Designated States: CA, JP, MX, European patent (AT, BE, CH, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE).

Published
With international search report.

(54) Title: HIGH CAPACITY WIRELESS COMMUNICATION USING SPATIAL SUBCHANNELS



(57) Abstract

In a system and method of digital wireless communication between a base station (B) and a subscriber unit (S), a spatial channel characterized by a channel matrix H couples an adaptive array of Mt antenna elements (1-Mt) at the base station (B) with an adaptive array Mr antenna elements (1-Mr) at the subscriber unit (S). The method comprises the step of determining from the channel matrix H a number L of independent spatio-temporal subchannels, and encoding a plurality of information signals into a sequence of transmitted signal vectors. The transmitted signal vectors have Mt complex valued components and are selected to transmit distinct signal information in parallel over the independent subchannels, thereby providing increased communication capacity between the base and the subscriber. The sequence of transmitted signal vectors is transmitted from the base station array (1-Mt), and a sequence of received signal vectors is received at the subscriber array (1-Mr) and are decoded to yield the original information signals. Specific spatio-temporal coding techniques are described that increase system performance.

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High Capacity Wireless Communication Using Spatial Subchannels

RELATED APPLICATIONS

This application claims priority from U.S. provisional applications 60/025,227 and 60/025,228, both filed 08/29/96. Both applications are hereby incorporated by reference.

FIELD OF THE INVENTION

This invention relates generally to digital wireless communication systems. More particularly, it relates to using antenna arrays by both a base station and a subscriber to significantly increase the capacity of wireless communication systems.

BACKGROUND OF THE INVENTION

Due to the increasing demand for wireless communication, it has become necessary to develop techniques for more efficiently using the allocated frequency bands, i.e. increasing the capacity to communicate information within a limited available bandwidth. This increased capacity can be used to enhance system performance by increasing the number of information channels, by increasing the channel information rates and/or by increasing the channel reliability.

FIG. 1 shows a conventional low capacity wireless communication system. Information is transmitted from a base station B to subscribers $S_1, \ldots, S_9$ by broadcasting omnidirectional signals on one of several predetermined frequency channels. Similarly, the subscribers transmit information back to the base station by broadcasting similar signals on one of the frequency channels. In this system, multiple users independently access the system through the

division of the frequency band into distinct subband frequency channels. This technique is known as frequency division multiple access (FDMA).

A standard technique used by commercial wireless phone systems to increasing capacity is to divide the service region into spatial cells, as shown in FIG. 2. Instead of using just one base station to serve all users in the region, a collection of base stations $B_1, \ldots, B_7$ are used to independently service separate spatial cells. In such a cellular system, multiple users can reuse the same frequency channel without interfering with each other, provided they access the system from different spatial cells. The cellular concept, therefore, is a simple type of spatial division multiple access (SDMA).

In the case of digital communication, additional techniques can be used to increase capacity. A few well known examples are time division multiple access (TDMA) and code division multiple access (CDMA). TDMA allows several users to share a single frequency channel by assigning their data to distinct time slots. CDMA is normally a spread-spectrum technique that does not limit individual signals to narrow frequency channels but spreads them throughout the frequency spectrum of the entire band. Signals sharing the band are distinguished by assigning them different orthogonal digital code sequences. These techniques use digital coding to make more efficient use of the available spectrum.

Wireless systems may also use combinations of the above techniques to increase capacity, e.g. FDMA/CDMA and TDMA/CDMA. Although these and other known techniques increase the capacity of wireless communication systems, there is still a need to further increase system performance. Recently, considerable attention has focused on ways to increasing capacity by further exploiting the spatial domain.

One well-known SDMA technique is to provide the base station with a set of independently controlled directional antennas, thereby dividing the cell into separate sectors, each controlled by a separate antenna. As a result, the frequency reuse in the system can be increased and/or cochannel interference can be reduced. Instead of independently controlled directional antennas, this technique can also be implemented with a coherently controlled antenna array, as shown in FIG. 3. Using a signal processor to control the relative phases of the signals applied to the antenna elements, predetermined beams can be formed in the directions of the separate sectors. Similar signal processing can be used to selectively receive signals only from within the distinct sectors.

In an environment containing a significant number of reflectors (such as buildings), a signal will often follow multiple paths. Because multipath reflections alter the signal directions, the cell space experiences angular mixing and can not be sharply divided into distinct sectors. Multipath can therefore cause cochannel interference between sectors, reducing the benefit of sectoring the cell. In addition, because the separate parts of such a multipath signal can arrive with different phases that destructively interfere, multipath can result in unpredictable signal fading.

In order to avoid the above problems with multipath, more sophisticated SDMA techniques have been proposed. For example, U.S. Pat. No. 5,471,647 and U.S. Pat. No. 5,634,199, both to Gerlach et al., and U.S. Pat. No. 5,592,490 to Barratt et al. disclose wireless communication systems that increase performance by exploiting the spatial domain. In the downlink, the base station determines the spatial channel of each subscriber and uses this channel information to adaptively control its antenna array to form customized beams, as shown in FIG. 4A. These beams transmit an information

signal x over multiple paths so that the signal x arrives to the subscriber with maximum strength. The beams can also be selected to direct nulls to other subscribers so that cochannel interference is reduced. In the uplink, as shown in FIG. 4B, the base station uses the channel information to spatially filter the received signals so that the transmitted signal x' is received with maximum sensitivity and distinguished from the signals transmitted by other subscribers. In this approach the same information signal follows several paths, providing increased spatial redundancy.

In the uplink, there are well known signal processing techniques for estimating the spatial channel from the signals received at the base station antenna array, e.g. by using *a priori* spatial or temporal structures present in the signal, or by blind adaptive estimation. If the uplink and downlink frequencies are the same, then the spatial channel for the downlink is directly related to the spatial channel for the uplink, and the base can use the known uplink channel information to perform transmit beamforming in the downlink. Because the spatial channel is frequency dependent and the uplink and downlink frequencies are often different, the base does not always have sufficient information to derive the downlink spatial channel information. One technique for obtaining downlink channel information is for the subscriber to periodically transmit test signals to the base on the downlink frequency rather than the uplink frequency. Another technique is for the base to transmit test signals and for the subscriber to feedback channel information to the base. If the spatial channel is quickly changing due to the relative movement of the base, the subscriber and/or reflectors in the environment, then the spatial channel must be updated frequently, placing a heavy demand on the system. One method to reduce the required feedback rates is to track only the subspace spanned by the time-averaged channel vector, rather than the instantaneous channel vector. Even with this

reduction, however, the required feedback rates are still a large fraction of the signal information rate.

Although these adaptive beamforming techniques require substantial signal processing and/or large feedback rates to determine the spatial channel in real time, these techniques have the advantage that they can navigate the complex spatial environment and avoid, to some extent, the problems introduced by multipath reflections. As a result, an increase in performance is enjoyed by adaptive antenna array systems, due to their use of the spatial dimension. Note, however, that while the base station antenna array can make efficient use of the spatial dimension by selectively directing the downlink signal to the subscriber S, the uplink signal in these systems is spatially inefficient. Typically, the subscriber is equipped with only a single antenna that radiates signal energy in all directions, potentially causing cochannel interference. These communication systems, therefore, do not make optimal use of the spatial dimension to increase capacity.

OBJECTS AND ADVANTAGES OF THE INVENTION

Accordingly, it is a primary object of the present invention to provide a communication system that significantly increases the capacity and performance of wireless communication systems by taking maximum advantage of the spatial domain. Another object of the invention is to provide computationally efficient coding techniques that make optimal use of the spatial dimensions of the channel. These and other objects and advantages will become apparent from the following description and associated drawings.

SUMMARY OF THE INVENTION

These objects and advantages are attained by a method of digital wireless communication that takes maximal advantage of spatial channel dimensions between a base station and a subscriber unit to increase system capacity and performance.

Surprisingly, the techniques of the present invention provide an increased information capacity in multipath environments. In contrast, known techniques suffer in the presence of multipath and do not exploit multipath to directly increase system capacity. In brief, the present invention teaches a method of wireless communication using antenna arrays at both the base and subscriber units to transmit distinct information signals over different spatial channels in parallel, thereby multiplying the capacity between the base and the subscriber. The present invention also teaches specific spatio-temporal coding techniques that make optimal use of these additional spatial subchannels.

Generally, the present invention provides a method of digital wireless communication between a base station and a subscriber unit, where a spatial channel characterized by a channel matrix **H** couples an array of M_T antenna elements at the base station with an array of M_R antenna elements at the subscriber unit. The method comprises the step of determining from the channel matrix **H** a number L of independent spatial subchannels, and encoding a plurality of information signals into a sequence of transmitted signal vectors. The transmitted signal vectors have M_T complex valued components and are selected to distribute distinct signal information over the independent spatial subchannels. The sequence of transmitted signal vectors is transmitted from the array of M_T antenna elements at the base station, and a sequence of received signal vectors is received at the array of M_R antenna elements at the subscriber unit. The received signal vectors have M_R complex valued components. These received signal vectors are decoded to yield the information signals.

In another aspect, the invention provides a method that comprises computing from a set of K original information signals a spatio-temporal coded signal in accordance with a channel matrix **H**. The channel matrix **H** represents the spatio-temporal characteristics of the information link between a

base station array of M_T antenna elements and a subscriber unit array of M_R antenna elements. Signal processing techniques are used to decompose **H** into K parallel spatio-temporal subchannels that can independently carry information signals between the base and subscriber units. After transmitting the spatio-temporal coded signal over the channel, it is decoded into a set of K received information signals that correspond to the K original information signals. In a preferred embodiment, the K parallel spatio-temporal subchannels are characterized by a set of K spatio-temporal transmission sequences that are derived from a decomposition of **H** into independent modes, and a set of K corresponding receive sequences. For example, the K spatio-temporal transmission sequences may be multiples of right singular vectors of **H**, and the receive sequences may be a matched set of K spatio-temporal filter sequences that are left singular vectors of **H**.

If L is the number of multipath components between the base station and the subscriber unit, then the number K of parallel spatio-temporal channels is not more than $(N+\nu) \times M_R$, not more than $N \times M_T$, and not more than $N \times L$, where $(N+\nu)$ is a maximum number of nonzero output samples transmitted for a block of N symbols. In a preferred embodiment, the original information signals comprise K blocks of N symbols, and the channel matrix **H** comprises $M_T \times M_R$ blocks of $N \times (N+\nu)$ channel matrices $\mathbf{H}_{ij}$.

In some applications of the present invention, the channel state information (CSI) may not be completely known, or may be expensive to compute. Accordingly, the present invention also provides a method for facilitating the efficient computation of the K received information signals from the transmitted spatio-temporal coded signal by adding cyclic prefixes to the coded signal prior to transmission.

DESCRIPTION OF THE FIGURES

FIG. 1 shows a low capacity wireless communication system well known in the prior art.

FIG. 2 illustrates a known technique of spatially dividing a service region into cells in order to increase system capacity.

FIG. 3 illustrates the use of beamforming with an antenna array to divide a cell into angular sectors, as is known in the art.

FIGS. 4A and 4B illustrate state-of-the-art techniques using adaptive antenna arrays for downlink and uplink beamforming, respectively.

FIGS. 5A and 5B show the parallel transmission of distinct information signals using spatial subchannels in downlink and uplink, respectively, as taught by the present invention.

FIGS. 6A and 6B are physical and schematic representations, respectively, of a communication channel for a system with multiple transmitting antennas and multiple receiving antennas, according to the present invention.

FIG. 7 is a block diagram of the system architecture for communicating information over a multiple-input-multiple-output spatial channel according to the present invention.

DETAILED DESCRIPTION

Although the following detailed description contains many specifics for the purposes of illustration, anyone of ordinary skill in the art will appreciate that many variations and alterations to the following details are within the scope of the invention. Accordingly, the following preferred embodiment of the invention is set forth without any loss of generality to, and without imposing limitations upon, the claimed invention.

As discussed above in relation to FIGS. 4A and 4B, prior art wireless systems employing an adaptive antenna array at the

base station are multiple-input-single-output (MISO) systems, i.e. the channel from the base to the subscriber is characterized by multiple inputs at the transmitting antenna array and a single output at the receiving subscriber antenna. Because these MISO systems can exploit some of the spatial channel, they have an increased capacity as compared to single-input-single-output (SISO) systems that are discussed above in relation to FIGS. 1 and 2. It should be noted that although the MISO systems disclosed in the prior art provide an increase in overall system capacity by spatially isolating separate subscribers from each other, these systems do not provide an increase in the capacity of information transmitted from the base to a single subscriber, or vice versa. As shown in FIGS. 4A and 4B, only one information signal is transmitted between the base and subscriber in both downlink and uplink of a MISO system. Even in the case where the subscriber is provided with an antenna array, the prior art suggests only that this capability would further reduce cochannel interference. Although the overall system capacity could be increased, this would not increase the capacity between the base and a single subscriber.

The present invention, in contrast, is a multiple-input-multiple-output (MIMO) wireless communication system that is distinguished by the fact that it increases the capacity of both uplink and downlink transmissions between a base and a subscriber through a novel use of additional spatial channel dimensions. The present inventors have recognized the possibility of exploiting multiple parallel spatial subchannels between a base station and a subscriber, thereby making use of additional spatial dimensions to increase the capacity of wireless communication. Surprisingly, this technique provides an increased information capacity and performance in multipath environments, a result that is in striking contrast with conventional wisdom.

FIGS. 5A and 5B illustrate a MIMO wireless communication system according to the present invention. As shown in FIG. 5A, a base station B uses adaptive antenna arrays and spatial processing to transmit distinct downlink signals x_1, x_2, x_3 through separate spatial subchannels to a subscriber unit S which uses an adaptive array and spatial processing to receive the separate signals. In a similar manner, the subscriber S uses an adaptive array to transmit distinct uplink signals x'_1, x'_2, x'_3 to the base B over the same spatial subchannels. As the multipath in the environment increases, the channel acquires a richer spatial structure that allows more subchannels to be used for increased capacity.

It is important to note that the simple assignment of the distinct signals to the distinct spatial paths in a one-to-one correspondence, as illustrated above, is only one possible way to exploit the additional capacity provided by the spatial subchannel structure. For example, coding techniques can be used to mix the signal information among the various paths. In addition, the present inventors have developed techniques for coupling these additional spatial dimensions to available temporal and/or frequency dimensions prior to transmission. Although such coupled spatio-temporal coding techniques are more subtle than direct spatial coding alone, they provide better system performance, as will be described in detail below.

In order to facilitate an understanding of the present invention and enable those skilled in the art to practice it, the following description includes a teaching of the general principles of the invention, as well as implementation details. First we develop a compact model for understanding frequency dispersive, spatially selective wireless MIMO channels. We then discuss their theoretical information capacity limits, and propose spatio-temporal coding structures that asymptotically achieve theoretical channel capacity. In particular, a spatio-temporal vector coding (STVC) structure

for burst transmission is disclosed, as well as a more practical, reduced complexity, discrete matrix multitone (DMMT) space-frequency coding structure. Both STVC and DMMT are shown to achieve the theoretical channel capacity as the burst duration increases.

In its preferred implementations, the present invention makes use of many techniques and devices well known in the art of adaptive antenna arrays systems and associated digital beamforming signal processing. These techniques and devices are described in detail in U.S. Pat. No. 5,471,647 and U.S. Pat. No. 5,634,199, both to Gerlach et al., and U.S. Pat. No. 5,592,490 to Barratt et al., which are all incorporated herein by reference. In addition, a comprehensive treatment of the present state of the art is given by John Livita and Titus Kwok-Yeung Lo in *Digital Beamforming in Wireless Communications* (Artech House Publishers, 1996). Accordingly, the following detailed description focuses upon the specific signal processing techniques which are required to enable those skilled in the art to practice the present invention.

Consider a communication channel for a system with M_T transmitting antennas at a base B and M_R receiving antennas at a subscriber S, as illustrated in FIGS. 6A and 6B. The channel input at a sample time k can be represented by an M_T dimensional column vector

$$\mathbf{z}(k) = [z_1(k), \ldots, z_{M_T}(k)]^T,$$

and the channel output and noise for sample k can be represented, respectively, by M_R dimensional column vectors

$$\mathbf{x}(k) = [x_1(k), \ldots, x_{M_R}(k)]^T,$$

and

$$\mathbf{n}(k) = [n_1(k), \ldots, n_{M_R}(k)]^T.$$

The communication over the channel $\mathbf{H}$ may then be expressed as a vector equation

$$\mathbf{x}(k) = \mathbf{H}\mathbf{z}(k) + \mathbf{n}(k),$$

where the MIMO channel matrix is

$$\mathbf{H} = \begin{pmatrix} h_{1,1} \cdots h_{1,M_T} \\ \vdots \quad\quad \vdots \\ h_{M_R,1} \cdots h_{M_R,M_T} \end{pmatrix}.$$

Each matrix element h_{ij} represents the SISO channel between the i^{th} receiver antenna and the j^{th} transmitter antenna. Due to the multipath structure of the spatial channel, orthogonal spatial subchannels can be determined by calculating the independent modes (e.g. eigenvectors) of the channel matrix $\mathbf{H}$. These spatial subchannels can then be used to transmit independent signals and increase the capacity of the communication link between the base B and the subscriber S. Because the multipath introduces time delays, however, a spatial decomposition alone will result in temporal mixing of the signals. It is more appropriate, therefore, to perform a more general spatio-temporal analysis of the channel.

Let $\{z_j(n)\}$ be a digital symbol sequence to be transmitted from the j^{th} antenna element, $g(t)$ a pulse shaping function impulse response, and T the symbol period. Then the signal applied to the j^{th} antenna element at time t is given by

$$s_j(t) = \sum_n z_j(n) g(t-nT)$$

The pulse shaping function is typically the convolution of two separate filters, one at the transmitter and one at the receiver. The optimum receiver filter is a matched filter. In practice, the pulse shape is windowed resulting in a finite duration impulse response. We assume synchronous complex baseband sampling with symbol period T. We define n_0 and $(\nu+1)$

to be the maximum lag and length over all paths l for the windowed pulse function sequences $\{g(nT - \tau_l)\}$. To simplify notation, it is assumed that $n_0 = 0$, and the discrete-time notation $g(nT - \tau_l) = g_l(n)$ is adopted.

When a block of N data symbols are transmitted, $N+\nu$ non-zero output samples result beginning at time sample $k-N+1$ and ending with sample $k+\nu$. The composite channel output can now be written as an $M_R \aleph (N+\nu)$ dimensional column vector with all time samples for a given receive antenna appearing in order so that

$$\mathbf{x}(k) = [x_1(k-N+1), \ldots, x_1(k+\nu), \ldots, x_{M_R}(k-N+1), \ldots, x_{M_R}(k+\nu)]^T,$$

with an identical stacking for the output noise samples $\mathbf{n}(k)$. Similarly, the channel input is an $M_T \aleph N$ dimensional column vector written as

$$\mathbf{z}(k) = [z_1(k-N+1), \ldots, z_1(k), \ldots, z_{M_T}(k-N+1), \ldots, z_{M_T}(k)]^T,$$

The spatio-temporal communication over the channel $\mathbf{H}$ may then be expressed as a vector equation

$$\mathbf{x}(k) = \mathbf{H}\mathbf{z}(k) + \mathbf{n}(k),$$

where the MIMO channel matrix

$$\mathbf{H} = \begin{pmatrix} \mathbf{H}_{1,1} & \cdots & \mathbf{H}_{1,M_T} \\ \vdots & & \vdots \\ \mathbf{H}_{M_R,1} & \cdots & \mathbf{H}_{M_R,M_T} \end{pmatrix}$$

is composed of SISO sub-blocks $\mathbf{H}_{ij}$ with each sub-block possessing the well known Toeplitz form.

We will now discuss the information capacity for the spatio-temporal channel developed above. The following analysis assumes that the noise $\mathbf{n}(k)$ is additive white Gaussian noise (AWGN) with covariance $\sigma^2\mathbf{I}$. Each channel use consists of an N

symbol burst transmission and the total average power radiated from all antennas and all time samples is constrained to less than a constant.

Write the singular value decomposition (SVD) of the channel matrix as $\mathbf{H}=\mathbf{V}_H\Lambda_H\mathbf{U}_H^*$, with the j^{th} singular value denoted $\lambda_{H,j}$. Write the spatio-temporal covariance matrix for $\mathbf{z}(k)$ as $\mathbf{R}_z$ with eigenvalue decomposition $\mathbf{R}_z=\mathbf{V}_Z\Lambda_Z\mathbf{U}_Z^*$, and eigenvalues $\lambda_{Z,j}$.

It can be demonstrated that the information capacity for the discrete-time spatio-temporal communication channel defined above is given by

$$C = \sum_{n=1}^{N\aleph M_T} \log \left(1 + \frac{\lambda_{Z,n}|\lambda_{H,n}|^2}{\sigma^2}\right),$$

where $\lambda_{Z,n}$ is given by the spatio-temporal water-filling solution. Motivated by this result, the inventors devised the following temporal vector coding technique. By appropriately selecting up to $N\aleph M_T$ spatio-temporal transmission vectors that are multiples of the right singular vectors of $\mathbf{H}$, and receiving with up to $N\aleph M_T$ matched spatio-temporal filter vectors that are the left singular vectors of $\mathbf{H}$, up to $N\aleph M_T$ parallel spatio-temporal subchannels are constructed for communicating information over the channel. Mathematically, this STVC channel is derived as follows. Substituting $\mathbf{H}=\mathbf{V}_H\Lambda_H\mathbf{U}_H^*$ into the original equation $\mathbf{x}(k)=\mathbf{Hz}(k)+\mathbf{n}(k)$ for the channel gives

$$\mathbf{x}(k) = \mathbf{V}_H\Lambda_H\mathbf{U}_H^*\mathbf{z}(k) + \mathbf{n}(k),$$

Left multiplication by $\mathbf{V}_H^*$ yields

$$\mathbf{V}_H^*\mathbf{x}(k) = \Lambda_H\mathbf{U}_H^*\mathbf{z}(k) + \mathbf{V}_H^*\mathbf{n}(k),$$

which yields the STVC channel when rewritten as

$$\hat{\mathbf{x}}(k) = \Lambda_H \hat{\mathbf{z}}(k) + \hat{\mathbf{n}}(k),$$

where $\hat{\mathbf{z}}(k)=\mathbf{U}_H^{*}\mathbf{z}(k)$, $\hat{\mathbf{x}}(k)=\mathbf{V}_H^{*}\mathbf{x}(k)$ and $\hat{\mathbf{n}}(k)=\mathbf{V}_H^{*}\mathbf{n}(k)$.

By analyzing the ranks of the above matrices, it can be demonstrated that the maximum number of finite amplitude parallel spatio-temporal channel dimensions, K, that can be created to communicate over the far field channel defined above is equal to min { $N\aleph L$, $(N+\nu)\aleph M_R$, $N\aleph M$ }, where L is the number of multipath components. Thus, multipath is an advantage in far-field MIMO channels. If the multipath is large (L 1), the capacity can be multiplied by adding antennas to both sides of the radio link. This capacity improvement occurs with no penalty in average radiated power or frequency bandwidth because the number of parallel channel dimensions is increased. In practice, an adaptive antenna array base station, such as that described by Barratt et al., is modified to implement the above vector coding scheme. In particular, a signal processor is designed to perform a spatio-temporal transform of information signals in accordance with the above equations so that they may be transmitted through the independent parallel subchannels and decoded by the subscriber.

The space-time vector coding solution described above requires a computation of the singular value decomposition of an $(N+\nu)\aleph M_R \times N\aleph M_T$ matrix. Since this computation can be complex, the present inventors have developed an optimal space-time communication structure that requires less computational complexity to implement. In particular, complexity can be reduced by using a coding structure similar to the discrete multi-tone (DMT) standard. DMT is in widespread use for wired SISO channels. DMT has also been applied to wired MISO channels, as described in U.S. Pat. No. 5,625,651 which is hereby incorporated by reference. The present inventors have generalized DMT to the MIMO case and

adapted it to wireless channels to obtain a novel space-frequency coding structure that results in a matrix of transmission and reception vector solutions for each discrete Fourier transform (DFT) frequency index. Because this new coding scheme has been generalized to MIMO channels characterized by a channel matrix, it is called discrete matrix multi-tone (DMMT).

In DMMT, N data symbols are again transmitted during each channel usage. However, a cyclic prefix is added to the data so that the last ν data symbols are transmitted from each antenna element prior to transmitting the full block of N symbols. By receiving only N time samples at the output of each antenna element, ignoring the first and last ν output samples, the MIMO channel submatrices $\hat{\mathbf{H}}_{ij}$ now appear as cyclic structures:

$$\hat{\mathbf{H}}_{i,j} = \begin{pmatrix} h(\nu) & \ldots & h(0) & 0 & 0 \ldots 0 & 0 & 0 & 0 \\ 0 & h(\nu) & & h(0) & 0 \ldots 0 & 0 & 0 & 0 \\ \vdots & & & & \vdots \quad \vdots & & & \vdots \\ 0 & & & & 0 \ldots 0 & & & 0 \\ 0 & \ldots & 0 & 0 & 0 \ldots 0 & h(\nu) & \ldots & h(0) \\ & & & \vdots & \vdots \quad \vdots & & & \vdots \\ h(\nu-1) & \ldots & h(0) & 0 & 0 \ldots 0 & \ldots & 0 & h(\nu) \end{pmatrix}$$

Given the cyclic SISO channel blocks, the channel matrix can be diagonalized with a relatively simple three step procedure. First post multiply $\hat{\mathbf{H}}$ with the $N\aleph M_T \times N\aleph M_T$ block diagonal inverse discrete Fourier transform (IDFT) matrix $\mathbf{F}^{*(M_T)}$ where each diagonal block is the unitary $N \times N$ IDFT matrix $\mathbf{F}^*$. The next step is to premultiply $\hat{\mathbf{H}}$ by a similar $N\aleph M_R \times N\aleph M_R$ block diagonal DFT matrix $\mathbf{F}^{(M_R)}$ where the diagonal submatrices $\mathbf{F}$ are $N \times N$ DFT matrices. With the well known result that the discrete Fourier transform basis vectors form the orthonormal singular vectors of the cyclic matrices $\hat{\mathbf{H}}_{ij}$, the new channel

matrix resulting from the IDFT post multiplication and the DFT premultiplication is

$$\mathbf{F}^{(M_R)}\hat{\mathbf{H}}\mathbf{F}^{\star(M_T)} = \begin{pmatrix} \Gamma_{1,1} & \cdots & \Gamma_{1,M_T} \\ \vdots & & \vdots \\ \Gamma_{M_R,1} & \cdots & \Gamma_{M_R,M_T} \end{pmatrix}$$

where $\Gamma_{i,j}$ is the diagonal matrix containing the singular values $\gamma_{i,j,n}$ of the cyclic channel submatrix $\hat{\mathbf{H}}_{ij}$. Premultiplication and postmultiplication by a permutation matrix $\mathbf{P}$ yields the block diagonal matrix

$$\mathbf{P}\mathbf{F}^{(M_R)}\hat{\mathbf{H}}\mathbf{F}^{\star(M_T)}\mathbf{P} = \begin{pmatrix} \mathbf{B}_1 & & 0 \\ & \ddots & \\ 0 & & \mathbf{B}_N \end{pmatrix}$$

where

$$\mathbf{B}_n = \begin{pmatrix} \gamma_{1,1,n} & \cdots & \gamma_{1,M_T,n} \\ \vdots & & \vdots \\ \gamma_{M_R,1,n} & \cdots & \gamma_{M_R,M_T,n} \end{pmatrix}$$

is the $M_R \times M_T$ space-frequency channel evaluated at DFT index n. Given the SVD of $\mathbf{B}_n = \mathbf{V}_{B,n}\Lambda_{B,n}\mathbf{U}^{\star}_{B,n}$, the diagonal DMMT channel matrix $\hat{\mathbf{H}}$ is finally obtained by post multiplying by $\mathbf{U}_B^{(M_T)}$ and premultiplying by $\mathbf{V}_B^{\star(M_R)}$ to obtain

$$\Lambda_{\hat{H}} = \mathbf{V}_B^{\star(M_R)}\mathbf{P}\mathbf{F}^{(M_R)}\hat{\mathbf{H}}\mathbf{F}^{\star(M_T)}\mathbf{P}\mathbf{U}_B^{(M_T)} = \begin{pmatrix} \Lambda_{\hat{H},1} & & 0 \\ & \ddots & \\ 0 & & \Lambda_{\hat{H},N} \end{pmatrix}$$

where $\mathbf{U}_B^{(M_T)}$ is block diagonal containing the right singular matrices of the $\mathbf{B}_n$ matrices, $\mathbf{V}_B^{\star(M_R)}$ is block diagonal containing the left singular matrices of the $\mathbf{B}_n$ matrices, and each of the diagonal submatrices $\Lambda_{\hat{H},n}$ contains the DMMT spatial sub-channel amplitudes, $\lambda_{\hat{H},n,j}$ for DFT bin n. The parallel channel DMMT equation is then

$$\hat{\mathbf{x}}(k) = \Lambda_{\hat{H}}\hat{\mathbf{z}}(k) + \hat{\mathbf{n}}(k),$$

where $\mathbf{z}(k)$ is the dimension $N \times M_T$ input symbol vector, $\hat{\mathbf{x}}(k)$ is the dimension $N \times M_R$ output symbol vector, and $\hat{\mathbf{n}}(k)$ is the dimension $N \times M_R$ equivalent output noise vector after the DFT and spatial orthogonalization operations are performed. A block diagram architecture that implements this DMMT space-frequency channel decomposition is presented in FIG. 7. The left portion of the diagram corresponds to the application of the operators $\mathbf{F}^{*(M_T)}\mathbf{P}\mathbf{U}_B^{(M_T)}$ on the signal $\hat{\mathbf{z}}(k)$. These operations are performed by a signal processor at the transmitter. The right portion of the diagram corresponds to the application of the operators $\mathbf{V}_B^{*(M_R)}\mathbf{P}\mathbf{F}^{(M_R)}$ on the received signals to recover a received information signal $\hat{\mathbf{x}}(k)$. These operations are performed by a signal processor at the receiver. The central matrix $\hat{\mathbf{H}}$ corresponds to the spatial channel itself. By construction, the signal processing operations result in a direct relationship between the received and transmitted information signals, as indicated by the fact that the matrix $\Lambda_{\hat{H}}$ in the parallel channel DMMT equation is diagonal.

This coding scheme significantly reduces the signal processing complexity required at the transmitter and receiver to diagonalize all space-time subchannels for each data block. In particular, this asymptotically optimal space-frequency MIMO DMMT information transmission technique has a complexity advantage of approximately N^2 as compared to the vector coding case. Moreover, since all of the matrix operations involved in creating the diagonal DMMT channel are invertible, the capacity of the DMMT channel is unchanged from that of the original cyclic sub-block matrix $\hat{\mathbf{H}}$. Thus, compared to STVC, the only capacity decrease for the DMMT space-time coding solution is due to the radiated power penalty required to transmit the cyclic prefix. This capacity penalty, however,

becomes small for large N. Thus, this new communications structure offers the advantage of very large increases in capacity without penalty in total average transmitted power or bandwidth.

In order to perform transmit beamforming, the base station signal processor computes spatio-temporal downlink subchannel information from downlink channel information fed back from the subscriber. The downlink signal information is then encoded in accordance with this computed downlink subchannel information. Similarly, the subscriber performs the same functions for the uplink channel using information fed back from the base. Because the present invention provides techniques for efficient channel estimation and increased channel capacity, the base and subscriber can both quickly estimate the channel and exchange channel information over the increased capacity channels, possibly at a rate slower than that of information data. As a result, both the base and subscriber can maintain a high degree of spatial resolution in transmit beamforming, thereby significantly reducing cochannel interference from other base stations or subscribers. As a result of this high degree of spatial discrimination in both transmission and reception, many more base stations and subscribers can share the same region of space while using the same frequency channel. Consequently, in addition to increasing the capacity of the channel between any two arrays, the present invention also increases system wide capacity by significantly reducing cochannel interference.

The teaching contained in this description can easily be extended to channels where the noise is not white but is highly structured as in the case of additive co-channel interference. In this case, large gains in cellular network capacity result from the ability to null interference at the receiver and the ability to constrain radiated interference power at the transmitter. These spatial coding techniques can also be applied to single frequency subchannel systems with

flat fading, conventional analog multicarrier transmission channels, or CDMA channels where each code delay can be decomposed into orthogonal subchannels provided that there is sub-chip multipath. The concepts of the present invention can also be applied to a more general class of channels where the antenna array is distributed over large distances and the propagation does not follow far field behavior. Finally, other communication media such as wire-line, acoustic media, and optical media will experience the same basic communication system benefits when spatio-temporal MIMO channel structures are employed. Thus, it will be clear to one skilled in the art that the above embodiment may be altered in many ways without departing from the scope of the invention. Accordingly, the scope of the invention should be determined by the following claims and their legal equivalents.

CLAIMS

What is claimed is:

1. A method of digital wireless communication between a base station and a subscriber unit, the method comprising:
determining from channel information a number L of independent spatial subchannels, wherein the channel information comprises spatial information relating to a spatial channel coupling an array of M_T antenna elements at the base station with an array of M_R antenna elements at the subscriber unit;
encoding a plurality of information signals into a sequence of transmitted signal vectors, wherein the transmitted signal vectors have M_T complex valued components and are selected to send distinct information signal over the independent spatial subchannels;
transmitting the sequence of transmitted signal vectors from the array of M_T antenna elements at the base station;
receiving a sequence of received signal vectors at the array of M_R antenna elements at the subscriber unit, wherein the received signal vectors have M_R complex valued components; and
decoding the received signal vectors to recover the information signals.

2. The method of claim 1 further comprising transmitting the channel information from the subscriber to the base.

3. The method of claim 1 wherein the channel information comprises a spatio-temporal channel matrix.

4. The method of claim 1 wherein the number L of independent spatial subchannels is equal to the number of multiple signal paths between the base and the subscriber.

5. The method of claim 1 wherein the encoding step comprises scaling the information signals by complex numbers,

permuting the scaled information signals and inverse Fourier transforming the permuted scaled information signals, and wherein the decoding step comprises Fourier transforming the received signals, permuting the Fourier transformed received signals, and scaling the permuted Fourier transformed received signals.

6. A method of digital wireless communication between a base station and a subscriber unit, the method comprising:

computing from a set of K original information signals a spatio-temporal coded signal in accordance with a channel matrix $\mathbf{H}$ having K parallel spatio-temporal subchannels;

transmitting the spatio-temporal coded signal from a base station array of M_T antenna elements through a channel corresponding to the channel matrix $\mathbf{H}$ to a subscriber unit array of M_R antenna elements; and

computing from the transmitted spatio-temporal coded signal a set of K received information signals.

7. The method of claim 6 wherein K is not more than $(N+\nu)\times M_R$, not more than $N\times M_T$, and not more than $N\times L$, where L is a maximum number of multipath components between the base station and the subscriber unit, and where $(N+\nu)$ is a maximum number of nonzero output samples transmitted for a block of N symbols.

8. The method of claim 6 wherein the original information signals comprise K blocks of N symbols, and the channel matrix $\mathbf{H}$ comprises $M_T\times M_R$ blocks of $N\times(N+\nu)$ channel matrices $\mathbf{H}_{ij}$, where $(N+\nu)$ is a maximum number of nonzero output samples transmitted for a block of N symbols.

9. The method of claim 6 wherein cyclic prefixes are added to the coded signal prior to the transmitting step, thereby facilitating the efficient computation of the K received information signals from the transmitted spatio-temporal coded signal.

10. The method of claim 6 wherein the K parallel spatio-temporal subchannels are characterized by a set of K spatio-temporal transmission sequences that are derived from a decomposition of **H** into independent modes.

11. The method of claim 6 wherein the K parallel spatio-temporal subchannels are characterized by a set of K spatio-temporal transmission sequences that are multiples of right singular vectors of **H**, and matched set of K spatio-temporal filter sequences that are left singular vectors of **H**.

12. A digital wireless communication system comprising:

a base station comprising a base station antenna array and a base station signal processor coupled to the base station antenna array;

a subscriber unit comprising a subscriber antenna array coupled through a wireless channel to the base station antenna array and a subscriber signal processor coupled to the subscriber antenna array;

wherein the base station signal processor computes spatio-temporal downlink subchannel information from downlink channel information received from the subscriber, and encodes downlink signal information in accordance with the computed downlink subchannel information; and

wherein the subscriber signal processor computes spatio-temporal uplink subchannel information from uplink channel information received from the base station, and encodes uplink signal information in accordance with the computed uplink subchannel information.

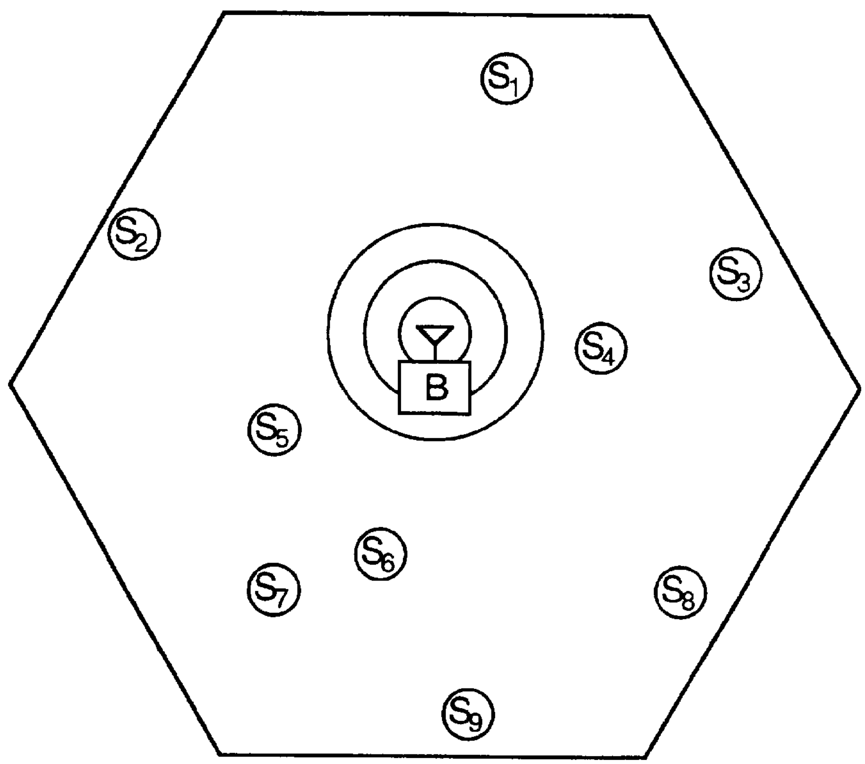


FIG. 1
(PRIOR ART)

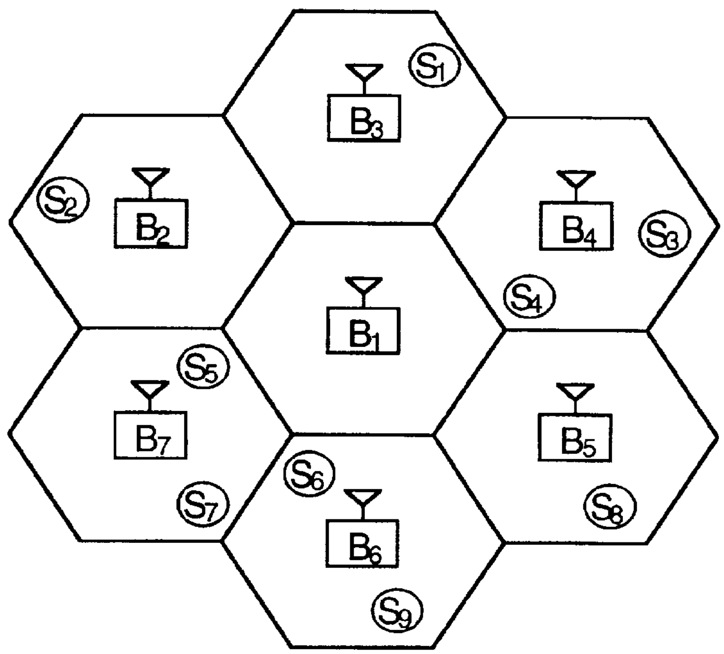


FIG. 2
(PRIOR ART)

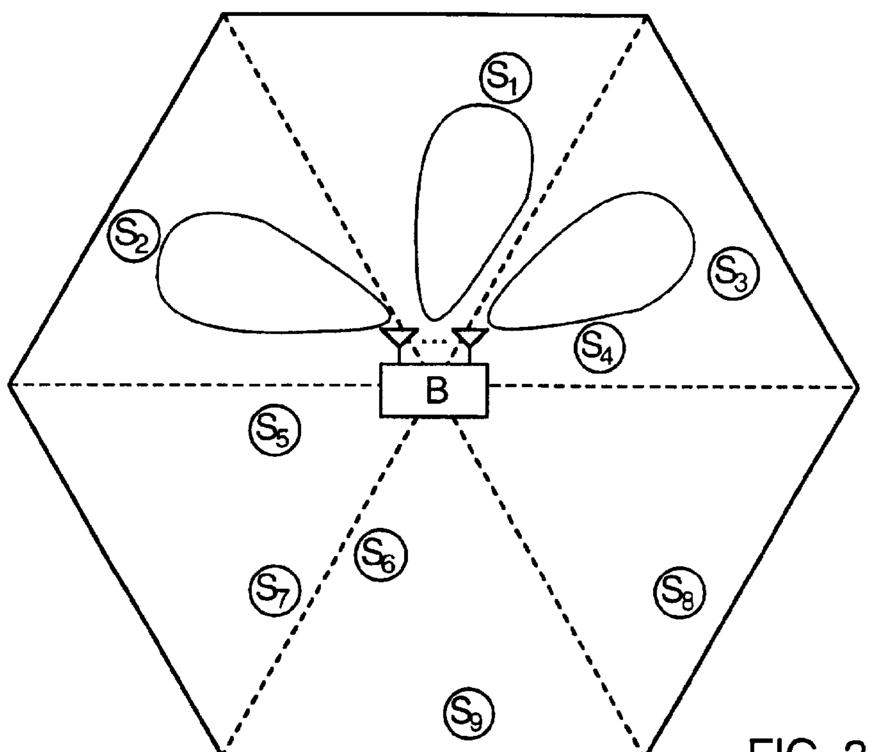


FIG. 3
(PRIOR ART)

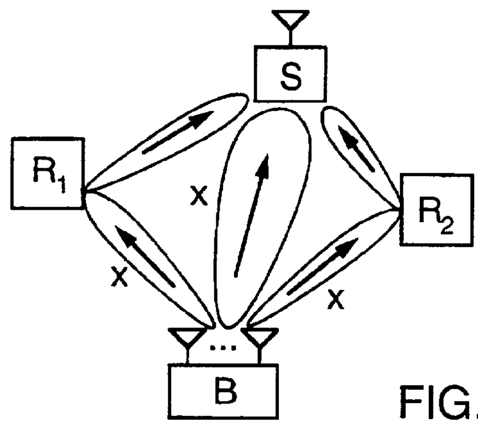


FIG. 4A
(PRIOR ART)

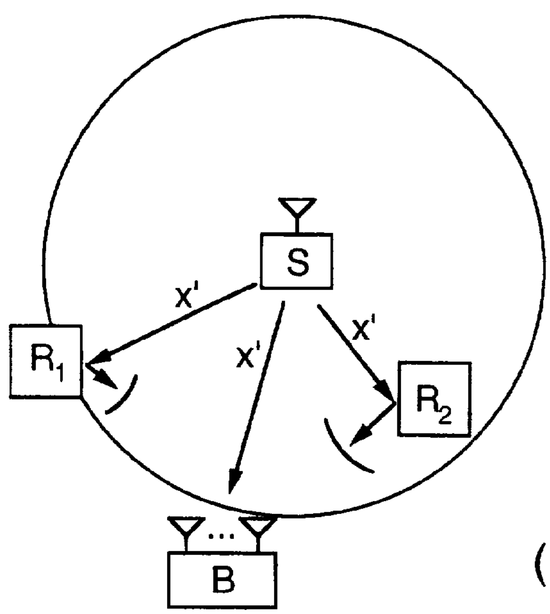


FIG. 4B
(PRIOR ART)

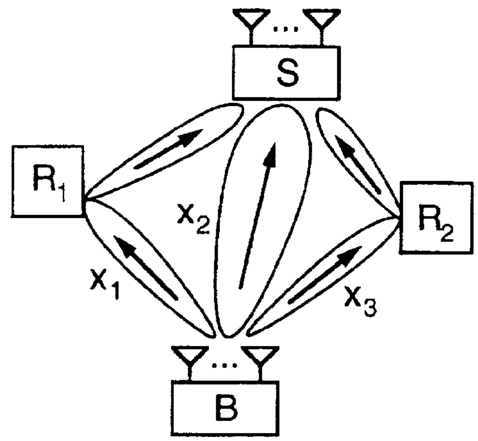


FIG. 5A

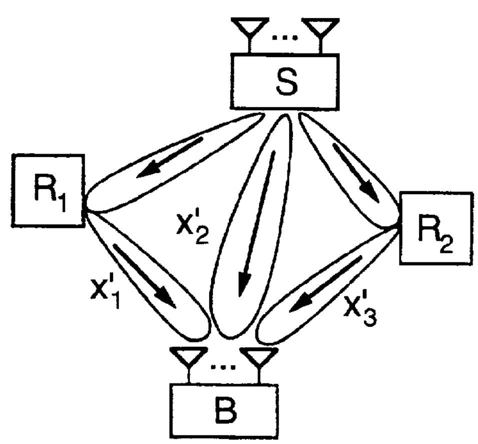
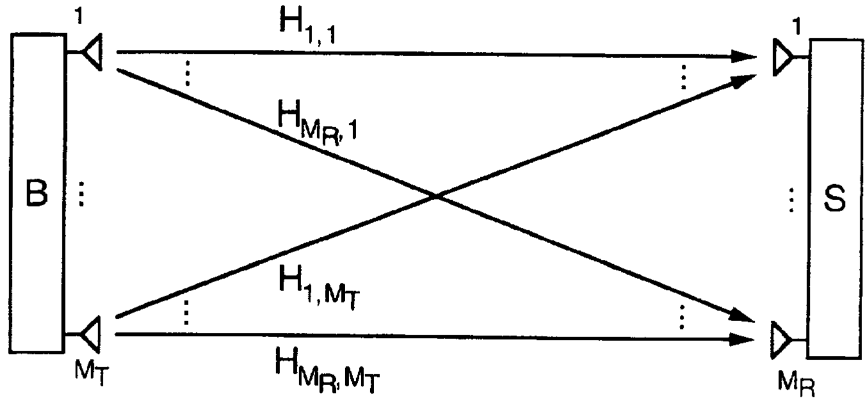
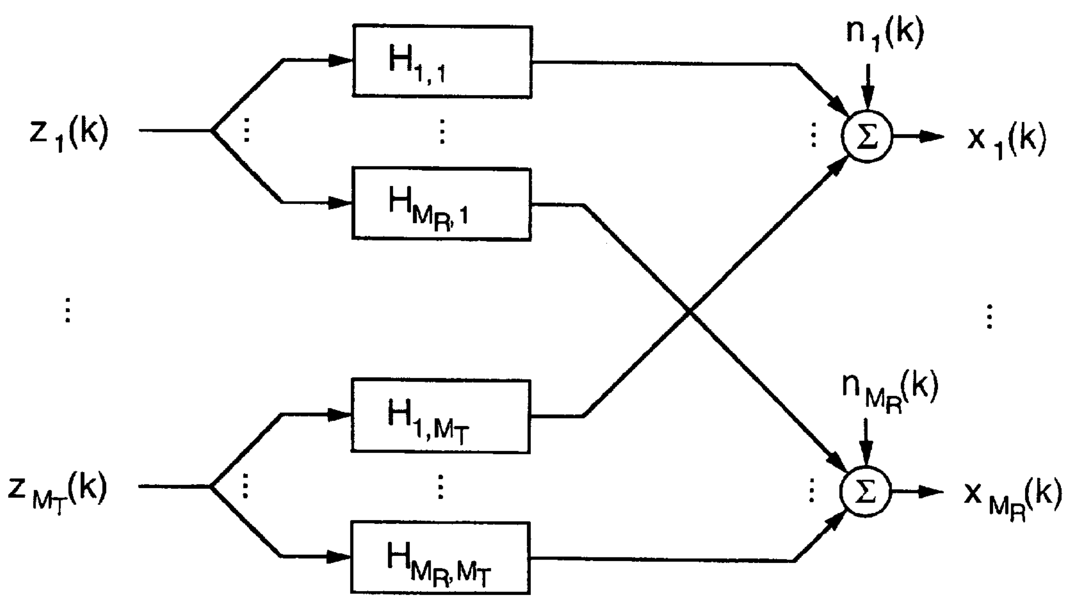


FIG. 5B



1
$H_{1,1}$
1
$H_{M_R,1}$
B
S
H_{1,M_T}
M_T
H_{M_R,M_T}
M_R

FIG. 6A



$n_1(k)$
$H_{1,1}$
$z_1(k)$
Σ
$x_1(k)$
$H_{M_R,1}$
$n_{M_R}(k)$
H_{1,M_T}
$z_{M_T}(k)$
Σ
$x_{M_R}(k)$
H_{M_R,M_T}

FIG. 6B

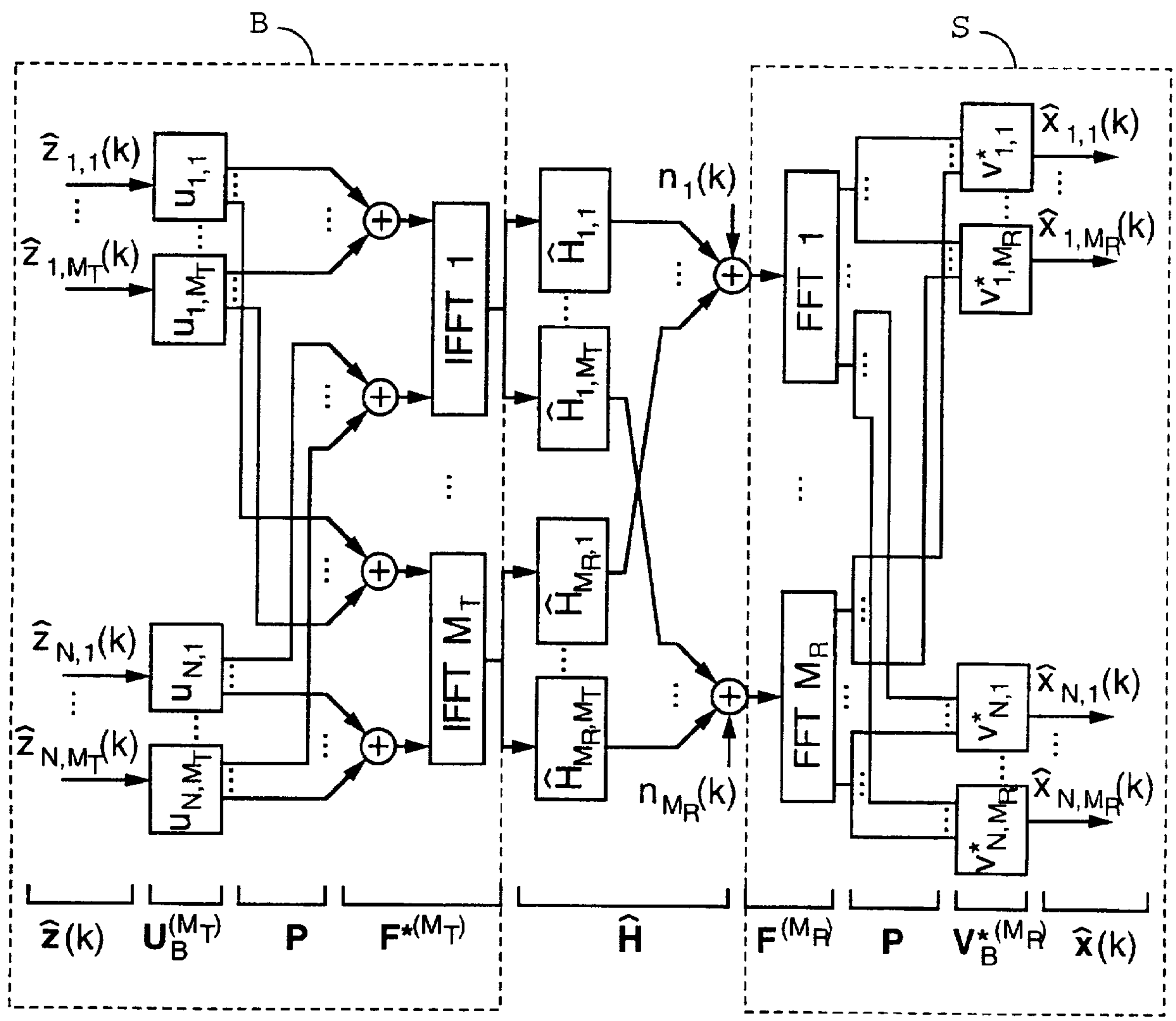

B
S
$\hat{z}_{1,1}(k)$
$\hat{z}_{1,M_T}(k)$
$\hat{z}_{N,1}(k)$
$\hat{z}_{N,M_T}(k)$
$u_{1,1}$
u_{1,M_T}
$u_{N,1}$
u_{N,M_T}
IFFT 1
IFFT M_T
$\hat{H}_{1,1}$
$\hat{H}_{1,M_T}$
$\hat{H}_{M_R,1}$
$\hat{H}_{M_R,M_T}$
$n_1(k)$
$n_{M_R}(k)$
FFT 1
FFT M_R
$v^*_{1,1}$
v^*_{1,M_R}
$v^*_{N,1}$
v^*_{N,M_R}
$\hat{x}_{1,1}(k)$
$\hat{x}_{1,M_R}(k)$
$\hat{x}_{N,1}(k)$
$\hat{x}_{N,M_R}(k)$
$\hat{\mathbf{z}}(k)$
$\mathbf{U}_B^{(M_T)}$
$\mathbf{P}$
$\mathbf{F}^{*(M_T)}$
$\hat{\mathbf{H}}$
$\mathbf{F}^{(M_R)}$
$\mathbf{P}$
$\mathbf{V}_B^{*(M_R)}$
$\hat{\mathbf{x}}(k)$

FIG. 7

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US97/15363

A. CLASSIFICATION OF SUBJECT MATTER

IPC(6) :H04B 1/38; H04M 1/00
US CL :455/562, 101, 103, 272
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 455/562, 101, 103, 272, 504, 506, 65, 132; 375/347

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched
NONE

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)
NONE

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 4,710,944 A (NOSSEN) 01 December 1987, columns 3-8, figures 1 and 5.	1-12
X	US 5,548,819 A (ROBB) 20 August 1996, columns 10-13, figures 1a-1b.	1-12
A,P	US 5,649,287 A (FORSSEN ET AL) 15 July 1997, figure 5.	1-12

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
09 OCTOBER 1997	17 NOV 1997

Name and mailing address of the ISA/US	Authorized officer
Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231	NGUYEN VO Jean Bell
Facsimile No. (703) 305-3230	Telephone No. (703) 308-6728

US006058105A

United States Patent [19]

Hochwald et al.

[11] **Patent Number: 6,058,105**

[45] **Date of Patent: May 2, 2000**

[54] **MULTIPLE ANTENNA COMMUNICATION SYSTEM AND METHOD THEREOF**

[75] Inventors: **Bertrand M. Hochwald**; **Thomas Louis Marzetta**, both of Summit, N.J.

[73] Assignee: **Lucent Technologies Inc.**, Murray Hill, N.J.

[21] Appl. No.: **08/938,168**

[22] Filed: **Sep. 26, 1997**

[51] **Int. Cl.**^{7} **H04B 1/02**; H04B 1/04; H04B 7/00

[52] **U.S. Cl.** **370/310**; 455/103; 342/367

[58] **Field of Search** 370/277, 280, 370/310, 326, 329, 334, 315, 316, 323, 325; 455/103, 132, 272, 562; 342/354, 367

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"Capacity of Multi–antenna Gaussian Channels," by I. Emre Telatar, *AT&T Bell Laboratories Technical Memorandum*, Document No. BL011217–950615–07TM, Nov. 13,1991.

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Primary Examiner—Michael Horabik
Assistant Examiner—Kevin C. Harper
Attorney, Agent, or Firm—Julio A. Garceran

[57] **ABSTRACT**

A communications system achieves high bit rates over an actual communications channel between M transmitter antennas of a first unit and N receiver antennas of a second unit, where M or N>1, by creating virtual sub-channels from the actual communications channel. The multiple antenna system creates the virtual sub-channels from the actual communications channel by using propagation information characterizing the actual communications channel at the first and second units. For transmissions from the first unit to the second unit, the first unit sends a virtual transmitted signal over at least a subset of the virtual sub-channels using at least a portion of the propagation information. The second unit retrieves a corresponding virtual received signal from the same set of virtual sub-channels using at least another portion of said propagation information.

36 Claims, 5 Drawing Sheets

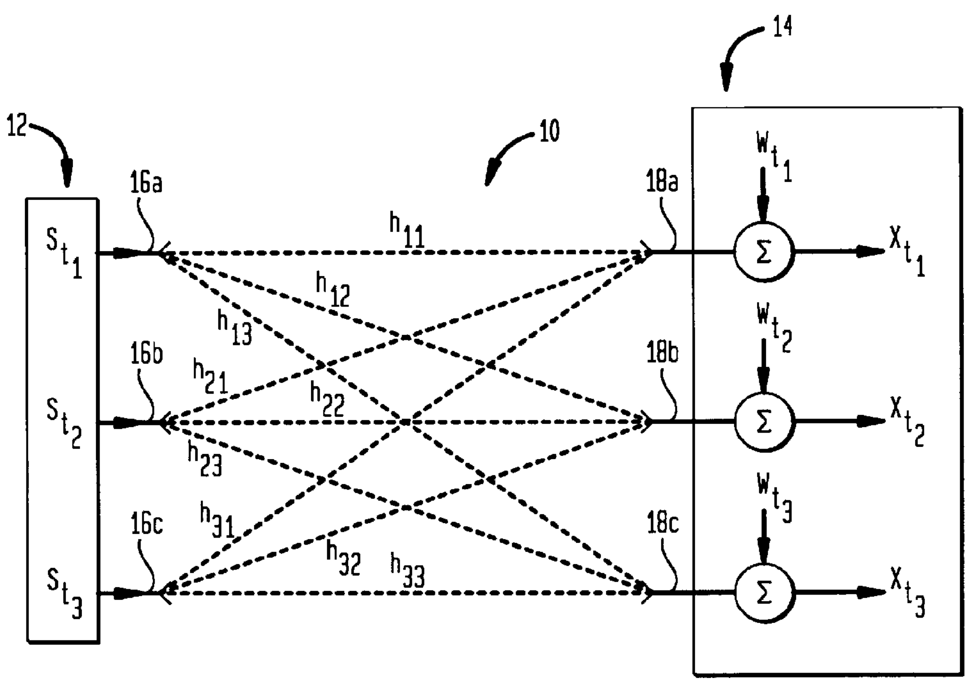


FIG. 1

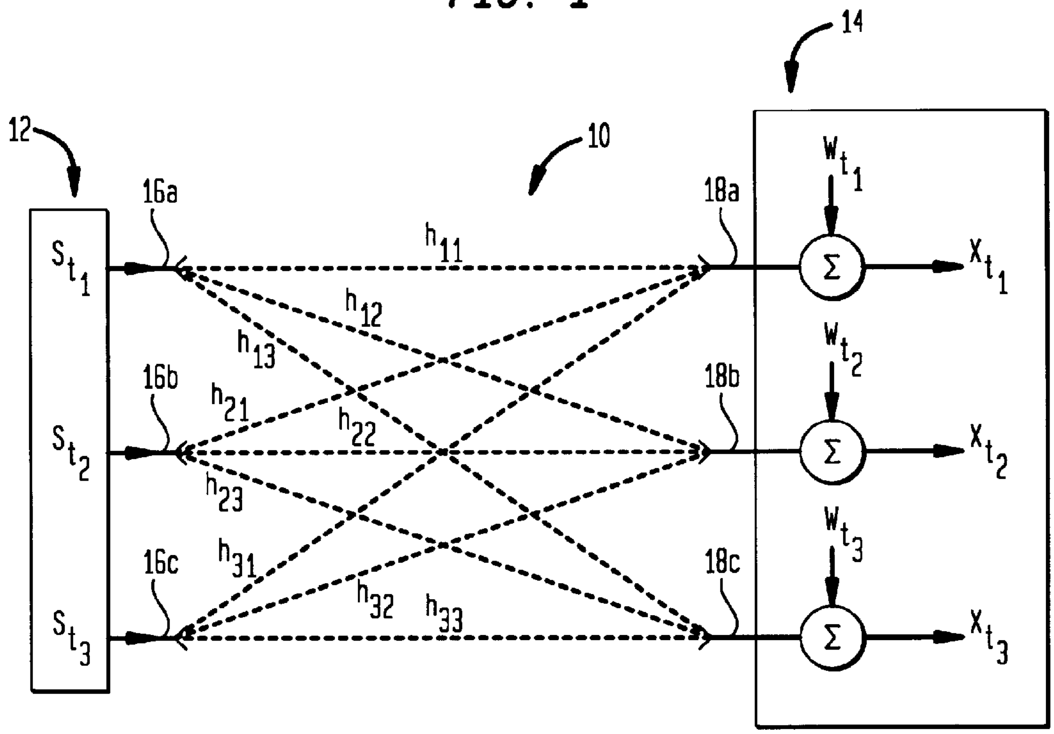

14
12
10
16a
18a
S_{t_1}
W_{t_1}
h_{11}
X_{t_1}
h_{12}
h_{13}
16b
18b
h_{21}
W_{t_2}
h_{22}
S_{t_2}
X_{t_2}
h_{23}
16c
h_{31}
18c
W_{t_3}
h_{32}
h_{33}
S_{t_3}
X_{t_3}

FIG. 2

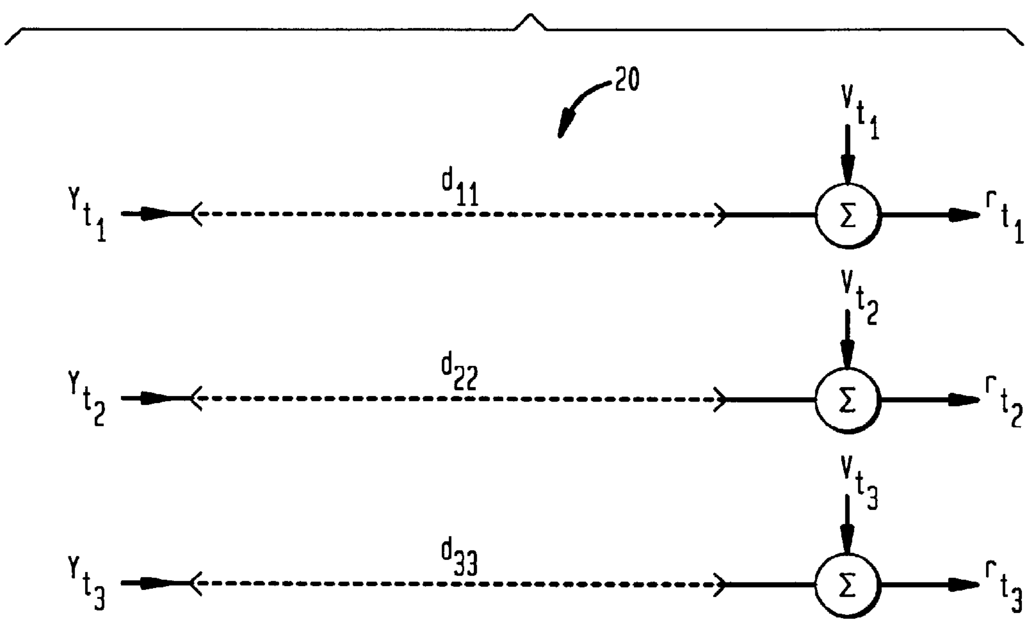
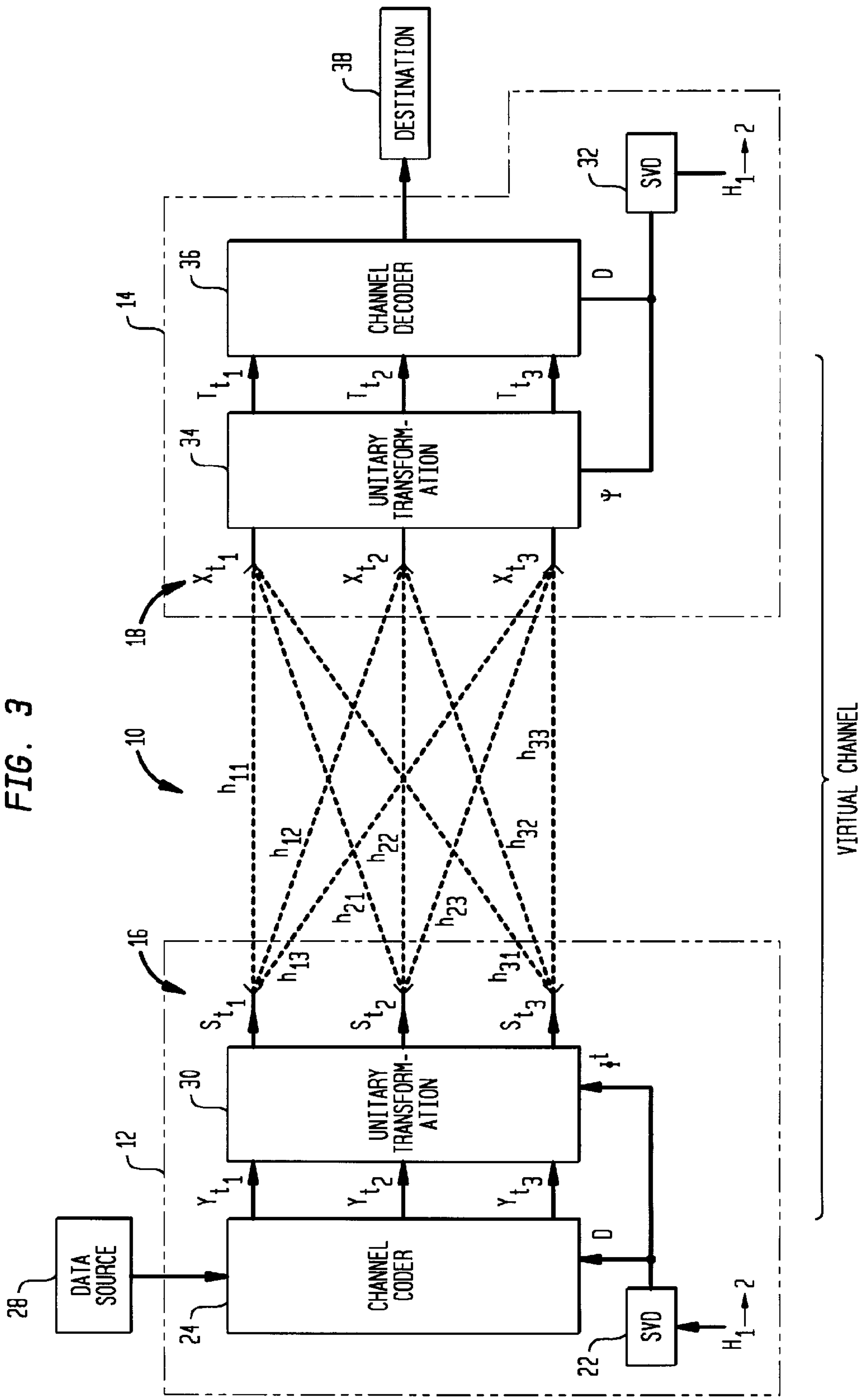

20
V_{t_1}
Y_{t_1}
d_{11}
r_{t_1}
V_{t_2}
Y_{t_2}
d_{22}
r_{t_2}
V_{t_3}
Y_{t_3}
d_{33}
r_{t_3}

FIG. 3


DESTINATION
38
SVD
32
$H_{1\to2}$
CHANNEL DECODER
36
14
D
T_{t_1}
T_{t_2}
T_{t_3}
UNITARY TRANSFORM-ATION
34
Ψ
X_{t_1}
X_{t_2}
X_{t_3}
18
10
h_{11}
h_{12}
h_{13}
h_{21}
h_{22}
h_{23}
h_{31}
h_{32}
h_{33}
16
S_{t_1}
S_{t_2}
S_{t_3}
UNITARY TRANSFORM-ATION
30
$\Phi^{\dagger}$
12
Y_{t_1}
Y_{t_2}
Y_{t_3}
DATA SOURCE
28
CHANNEL CODER
24
D
SVD
22
$H_{1\to2}$
VIRTUAL CHANNEL

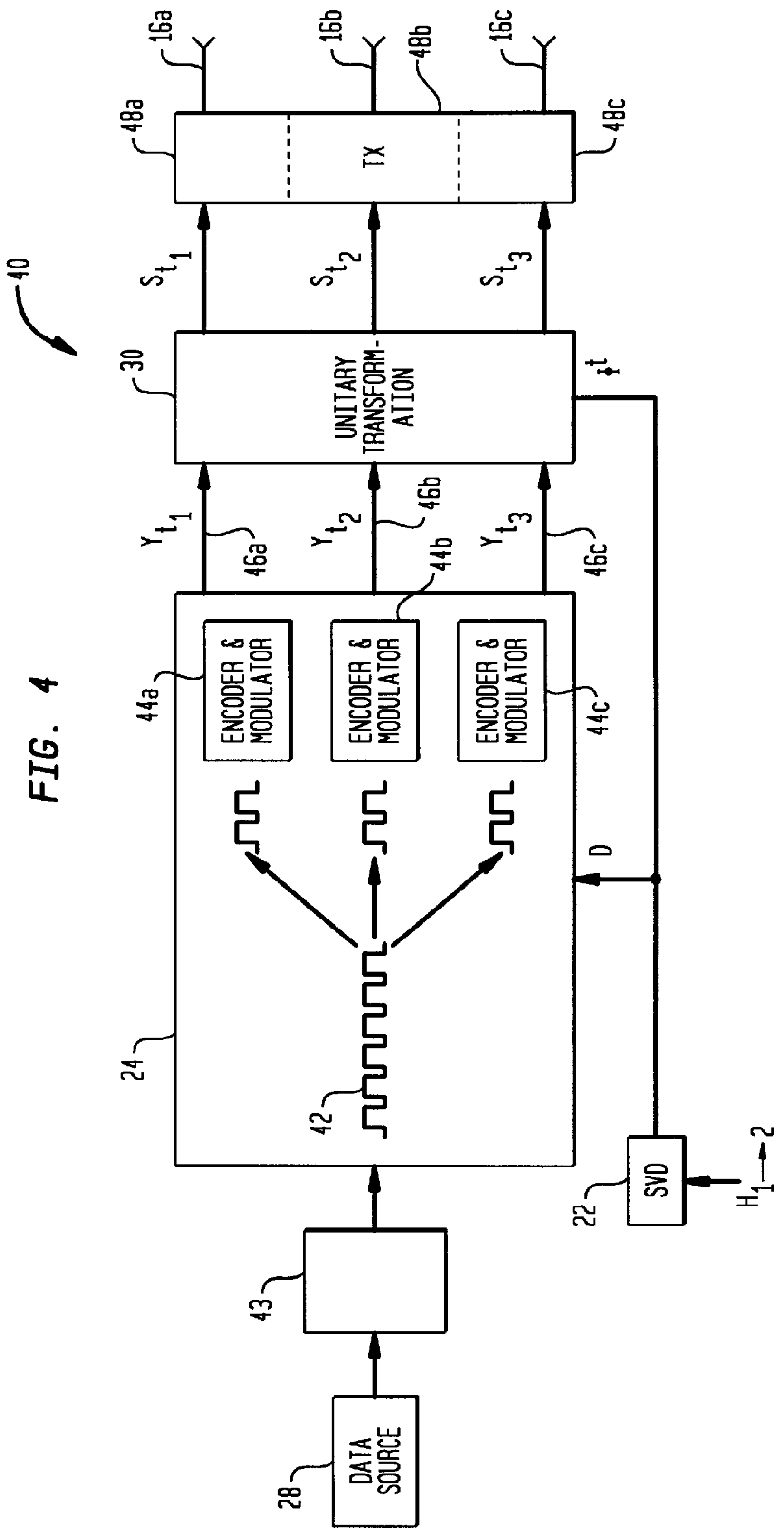

40
16a
16b
16c
48a
48b
48c
TX
S_{t_1}
S_{t_2}
S_{t_3}
30
UNITARY TRANSFORM-ATION
t
Y_{t_1}
Y_{t_2}
Y_{t_3}
46a
46b
46c
44a
44b
44c
ENCODER & MODULATOR
ENCODER & MODULATOR
ENCODER & MODULATOR
D
24
42
43
22
SVD
$H_{1 \to 2}$
28
DATA SOURCE

FIG. 4

FIG. 5

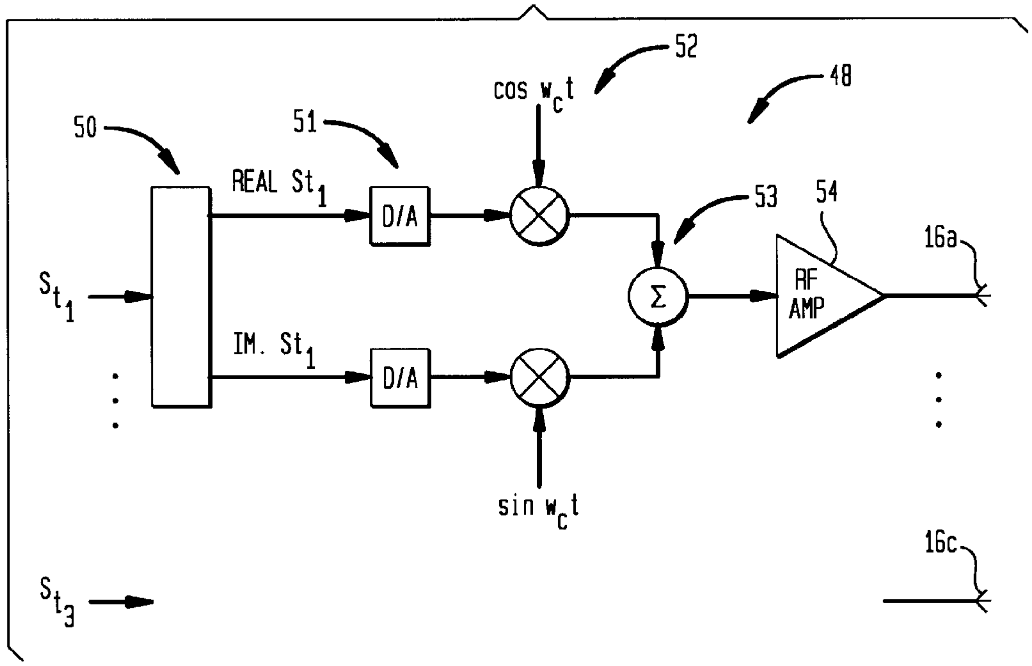
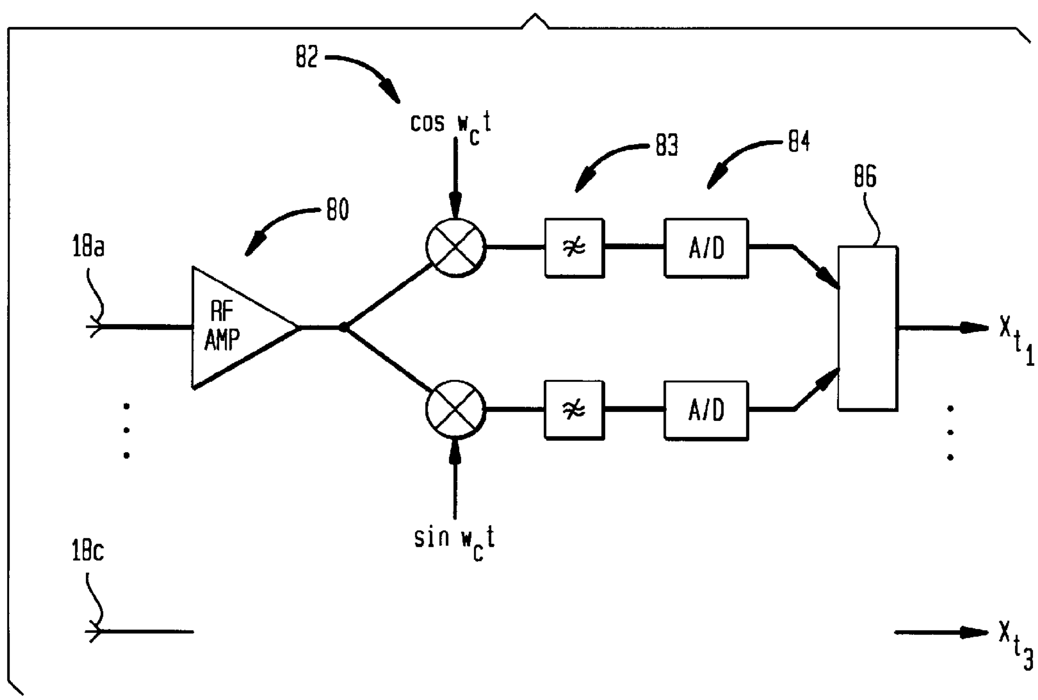

52
48
cos $w_c t$
50
51
REAL St_1
D/A
53
54
S_{t_1}
Σ
RF AMP
16a
IM. St_1
D/A
sin $w_c t$
16c
S_{t_3}

FIG. 7


82
cos $w_c t$
83
84
86
80
18a
RF AMP
A/D
A/D
X_{t_1}
sin $w_c t$
18c
X_{t_3}

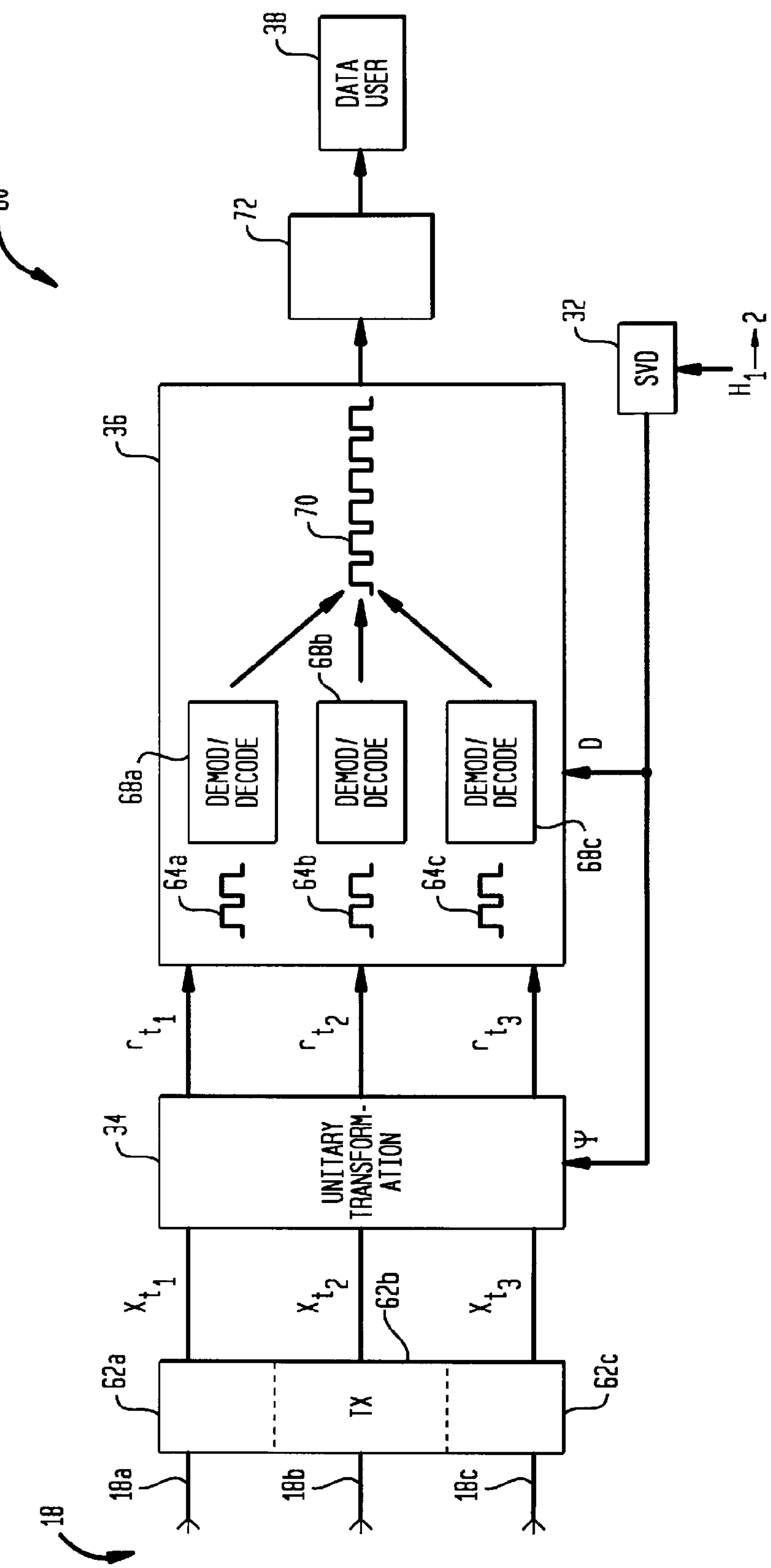

60
38
DATA USER
72
36
70
68b
68a
DEMOD/ DECODE
DEMOD/ DECODE
DEMOD/ DECODE
68c
D
64a
64b
64c
32
SVD
$H_{1\to 2}$
r_{t_1}
r_{t_2}
r_{t_3}
34
UNITARY TRANSFORM- ATION
Ψ
X_{t_1}
X_{t_2}
X_{t_3}
62a
62b
62c
TX
18
18a
18b
18c

FIG. 6

MULTIPLE ANTENNA COMMUNICATION SYSTEM AND METHOD THEREOF

BACKGROUND OF THE INVENTION

1. Field of the Invention

This invention relates to wireless communication systems and, more particularly, to a multiple antenna communication system.

2. Description of Related Art

The ultimate bit rate at which a digital wireless communication system may communicate data may be derived using Shannon's approach to information theory (commonly referred to as the Shannon limit). The ultimate bit rate is based on a number of different parameters, including: the total radiated power at the transmitter; the number of antennas at the transmitter and receiver; bandwidth; noise power at the receiver; and the characteristics of the propagation environment. For wireless transmission using multiple antennas at the transmitter and/or receiver in a so-called Rayleigh fading environment, the ultimate bit rate could be enormous, for example, hundreds of bits per second per Hz for a system employing 30 antennas at both the transmitter and receiver and experiencing an average signal-to-noise ratio of 16 dB.

A need exists for a wireless communications system which achieves high bit rates in a cost effective and relatively simple manner.

SUMMARY OF THE INVENTION

The present invention involves a communications system that achieves high bit rates over an actual communications channel between M transmitter antennas of a first unit and N receiver antennas of a second unit, where M or N>1, by creating virtual sub-channels from the actual communications channel. The multiple antenna system creates the virtual sub-channels from the actual communications channel by using at the first and second units propagation information characterizing the actual communications channel. For transmissions from the first unit to the second unit, the first unit sends a virtual transmitted signal over at least a subset of the virtual sub-channels using at least a portion of the propagation information. The second unit retrieves a corresponding virtual received signal from the same set of virtual sub-channels using at least another portion of said propagation information.

In general, a propagation matrix of propagation coefficients characterizes the propagation of communication signals between the transmitting antenna(s) of the first unit and the receiving antenna(s) of the second unit. By knowing the propagation characteristics of the actual communications channel (multiple-antenna channel), the multiple antenna system can decompose the actual communications channel into multiple virtual sub-channels. For transmissions from the first unit to the second unit, both the first unit and the second unit obtain propagation information which characterizes the transmissions from the first unit to the second unit. In certain embodiments, the first unit obtains at least a portion of the propagation information, and the second unit obtains at least another portion of the propagation information. Using the respective portions of the propagation information, the first and second units cooperatively render the actual communications channel into virtual sub-channels, thereby achieving high bit rate or throughput in a relatively simple manner.

In certain embodiments, the first and second units obtain the propagation matrix as the propagation information for transmissions from the first unit to the second unit. Initially, the first unit and second units obtain the propagation matrix by an exchange of signals. For example, the first unit transmits training signals to the second unit. From the training signals as transmitted and the training signals as received over the actual communications channel, the propagation matrix can be determined. Once the propagation matrix is determined, each unit can perform a singular value decomposition of the propagation matrix. The singular value decomposition of the propagation matrix yields the propagation matrix as the product of three factors D, Φ and ψ^+, where D is a diagonal matrix and Φ and ψ^+ are two unitary matrices with the superscript "+" denoting a conjugate transpose. The singular value decomposition serves to diagonalize the propagation matrix. The number of nonzero diagonal elements in the diagonal matrix D corresponds to the number of parallel independent virtual sub-channels for the actual communications channel. In some embodiments, for transmissions from the first unit to the second unit, the first unit obtains at least a portion of the propagation information which includes the diagonal matrix D and the unitary matrix Φ. The first unit provides the diagonal matrix D to a channel coder/modulator to encode and modulate an incoming bit or information stream onto the independent virtual sub-channels according to the values of the diagonal matrix D to produce a virtual transmitted signal. As such the diagonal matrix D can provide relative scaling of the bit rate. The first unit then performs a unitary transformation on the virtual transmitted signal by multiplying the virtual transmitted signal with the conjugate transpose of the unitary matrix Φ to produce the actual transmitted signal.

The second unit obtains at least another portion of the propagation information which includes the unitary matrix ψ^+ and the diagonal matrix D in certain embodiments. The second unit performs a unitary transformation on the actual received signal by multiplying the actual received signal with the unitary matrix ψ to produce a virtual received signal. The multiplications at the first and second units by the unitary matrices establish a virtual channel from the actual communications channel between the virtual transmitted signal and the virtual received signal which can be treated as parallel independent virtual sub-channels. The second unit provides the diagonal matrix D to a channel decoder/demodulator to decode and demodulate the virtual received signal according to the matrix D to produce an information stream. Thus, the multiple antenna system provides high capacity by effectively providing parallel independent sub-channels within the same frequency band. The multiple antenna system also provides enhanced performance because the multiple antenna system transmits bits on the virtual sub-channels relative to the values of the diagonal matrix D, thereby the stronger virtual sub-channels are used to transmit more information.

BRIEF DESCRIPTION OF THE DRAWINGS

Other aspects and advantages of the present invention may become apparent upon reading the following detailed description and upon reference to the drawings in which:

FIG. **1** shows a baseband representation of a multiple antenna channel or the actual communications channel used according to the principles of the present invention;

FIG. **2** shows a baseband representation of the virtual sub-channels derived from the actual communications channel according to certain principles of the present invention;

FIG. **3** shows a block diagram of an embodiment of the multiple antenna system according to the principles of the present invention;

FIG. **4** shows a block diagram of an embodiment of a multiple antenna transmitter according to certain principles of the present invention; and

FIG. **5** shows a block diagram of a particular embodiment of the transmit circuitry for the transmitter of FIG. **4**;

FIG. **6** shows a block diagram of an embodiment of a multiple antenna receiver according to certain principles of the present invention; and

FIG. **7** shows a block diagram of an embodiment of the receive circuitry of the receiver of FIG. **6**.

DETAILED DESCRIPTION OF THE DRAWINGS

An illustrative embodiment of the multiple antenna communication system according to the principles of the present invention is described below as the multiple antenna communication system might be implemented to provide high bit rate and enhanced performance. The multiple antenna system accomplishes this by using multiple antenna arrays at the transmitter and/or receiver and taking advantage of the propagation characteristics obtained for the multiple-antenna channel between the antenna(s) of a first unit and the antenna(s) of a second unit. By knowing certain propagation characteristics of the actual communications channel (multiple-antenna channel) at the first unit and the second unit, the multiple antenna system can achieve high bit rates by having the first unit and second unit cooperatively decompose the actual communications channel into multiple virtual sub-channels. For transmissions from the first unit to the second unit, the first and second units obtain at least respective portions of the propagation information characterizing the transmissions from the first unit to the second unit. The first and second units use at least their respective portions of the propagation information to decompose the actual communications channel into multiple virtual sub-channels over which communication signals are transmitted. As such, the multiple antenna communication system achieves high bit rates in a relatively simple manner without increasing total power or bandwidth by using the virtual sub-channels within the same frequency band. Additionally, the multiple antenna system provides enhanced performance by transmitting more bits over the stronger sub-channels as determined by the propagation information.

FIG. **1** shows a baseband representation for an actual communications channel **10** over which a first unit **12** transmits the RF signals corresponding to the components s_{t1}, s_{t2} and s_{t3} of an actual transmitted signal on carriers of the same frequency over respective multiple antennas **16***a–c* to a second unit **14**. In this particular embodiment, the first unit **12** has three antennas **16***a–c*, and the second unit **14** has three antennas **18***a–c*. The second unit **14** receives the RF signals at the respective receive antennas **18***a–c* corresponding to the components x_{t1}, x_{t2} and x_{t3} of the actual received signal. Each receive antenna **18***a–c* responds to each transmit antenna **16***a–c* through a complex-valued, scalar propagation coefficient h_{mn}, where m designates the respective transmit antenna **16***a–c* and n designates the respective receive antenna **18***a–c*. As such, the actual received signal x_{t1}, x_{t2} and x_{t3} can be characterized by the following:

$$x_{t1}=h_{11}\times s_{t1}+h_{21}\times s_{t2}+h_{31}\times s_{t3}+w_{t1}$$

$$x_{t2}=h_{12}\times s_{t1}+h_{22}\times s_{t2}+h_{32}\times s_{t3}+w_{t2}$$

$$x_{t3}=h_{13}\times s_{t1}+h_{23}\times s_{t2}+h_{33}\times s_{t3}+w_{t3},$$

where $\{w_{t1}, w_{t2}, w_{t3}\}$ are receiver noise added at the second unit **14**. In vector notation, $\underline{x}_t=\underline{s}_t\times H+\underline{w}_t$ where $\underline{x}_t=[x_{t1}, x_{t2}, x_{t3}]$, $\underline{s}_t=[s_{t1}, s_{t2}, s_{t3}]$, $\underline{w}_t=[w_{t1}, w_{t2}, w_{t3}]$, and the propagation matrix can be represented as:

$$H=\begin{matrix} h11 & h12 & h13 \\ h21 & h22 & h23 \\ h31 & h32 & h33 \end{matrix}.$$

In this particular embodiment, the second unit **14** can transmit signals in the reverse direction to the first unit **12**. The second unit **14** can use its receive antennas **18***a–c* to transmit the signals, and the first unit **12** can use its transmit antennas **16***a–c* to receive the signals. If the same frequency is used in both directions (using a time division duplex (TDD) scheme in certain bi-directional communications embodiments), the propagation coefficients for transmissions from the second unit **14** to the first unit **12** can be considered equal to the propagation coefficients for transmissions from the first unit **12** to the second unit **14** (according to the reciprocity principle), so $H_{2to1}=H^T_{1to2}$, where the superscript "T" denotes the transpose. If different frequencies are used in both directions (using a frequency division duplex scheme), the propagation coefficients for the signals propagating in both directions are determined. Using an appropriate training and set-up scheme, the first unit **12** and the second unit **14** can both learn the first unit **12** to second unit **14** propagation coefficients and the second unit **14** to first unit **12** propagation coefficients. For example, the units **12** and **14** could transmit known training signals over one antenna at a time and/or transmit training signals using orthogonal signals transmitted simultaneously over all respective antennas.

For ease of discussion, the multiple antenna scheme according to the principles of the present invention is discussed with particular reference to transmissions from the first unit **12** to the second unit **14**. It should be understood that the multiple antenna scheme can be applied to both uni-directional or bi-directional communications. In this particular embodiment, both units **12** and **14** perform a singular value decomposition (SVD) of the 3×3 propagation matrix H obtained by both units **12** and **14**. In this particular embodiment, the SVD of H yields $H=\Phi\times D\times\psi^+$, where D is a 3×3 real-valued, nonnegative, diagonal matrix,

$$D=\begin{matrix} d11 & 0 & 0 \\ 0 & d22 & 0 \\ 0 & 0 & d33 \end{matrix},$$

and Φ and ψ^+ are 3×3 complex unitary matrices with the superscript "+" denoting the "conjugate transpose." The columns of a unitary matrix have unit length and are mutually orthogonal. Multiplying a vector by a unitary matrix does not change the length of the vector but merely changes the direction of the vector. The inverse of a unitary matrix is equal to the conjugate transpose of the matrix, for example, $\Phi^+\times\Phi=I$ or for l=1 to 3, $\Sigma\Phi_{1i}{}^*\times\Phi_{1j}$ is equal to 1 if i=j and is equal to 0, if i≠j, where the superscript "*" denotes the "complex-conjugate."

The second unit **14** uses the matrix ψ to multiply the received signal $\underline{x}_t$ by the 3×3 unitary matrix ψ to obtain $\underline{r}_t=[r_{t1}, r_{t2}, r_{t3}]=\underline{x}_t\times\psi$. The first unit **12** uses the matrix Φ to transform the virtual transmitted signal $\underline{y}_t$ into the actual transmitted signal $\underline{s}_t$ by letting the actual transmitted signal $\underline{s}_t$ be equal to the conjugate transpose of the 3×3 unitary matrix Φ times the virtual transmitted signal $\underline{y}_t$, $\underline{s}_t=\underline{y}_t\times\Phi^+$ where $\underline{y}_t=[y_{t1}, y_{t2}, y_{t3}]$. The multiplications by the unitary matrices are invertible operations, so there is no loss of

information. In effect, a link is established between the virtual transmitted signal $\underline{y}_t$ and the virtual received signal $\underline{r}_t$, where

$$\begin{aligned}\underline{r}_t = \underline{x}_t \times \psi &= (\underline{s}_t \times H + \underline{w}_t) \times \psi \\ &= (\underline{y}_t \times \Phi^\dagger \times H + \underline{w}_t) \times \psi \\ &= \underline{y}_t \times \Phi^\dagger \times H \times \psi + \underline{w}_t \times \psi\end{aligned}.$$

Substituting $H=\Phi\times D\times\psi^+$ in the above expression gives:

$$\begin{aligned}\underline{r}_t &= \underline{y}_t \times \Phi^\dagger \times (\Phi \times D \times \psi^\dagger)\psi + \underline{w}_t \times \psi \\ &= \underline{y}_t \times D + \underline{v}_t, \text{ where } \underline{v}_t = \underline{w}_t \times \psi\end{aligned}.$$

Thus, the original link **10** between $\underline{s}_t$ and $\underline{x}_t$ is equivalent to a much simpler virtual link **20** between $\underline{y}_t$ and $\underline{r}_t$ as shown in FIG. **2**. In the actual communications channel, every one of the N receiver antennas responds to every one of the M transmitter antennas. The advantage of the virtual link or channel **10** is that it comprises virtual sub-channels, wherein each of a plurality of virtual receiver antennas responds to exactly one respective virtual transmitter antenna. In effect, the cross-coupled actual communications channel **10** of FIG. **1** can be treated as the three parallel, independent virtual sub-channels d_{11}, d_{22} and d_{33} shown in FIG. **2** according to the principles of the present invention.

The first and second units **12** and **14** can determine the respective propagation information needed to establish the virtual sub-channels between the first unit **12** and the second unit **14** in alternative ways. For example, since the matrix $HH^+=\Phi\times D\times D^+\times\Phi^+$, the first unit **12** and/or the second unit **14** can determine the unitary matrix Φ^+ from the eigenvectors of the matrix HH^+ and the diagonal matrix D from the eigenvalues of HH^+. Additionally, since $H^+H=\psi\times D^+\times D\times\psi^+$, the first unit **12** and/or the second unit **14** can determine the unitary matrix ψ from the eigenvectors of H^+H and the diagonal matrix D from the eigenvalues of H^+H. The diagonal matrix can be determined by using the squareroots of the eigenvalues of HH^+ or H^+H.

FIG. **3** shows an embodiment of the multiple antenna communication system according to certain principles of the present invention for transmissions between the first unit **12** and the second unit **14**. In this particular embodiment, the first unit **12** is shown as transmitting, and the second unit **14** is shown as receiving. For transmissions in the reverse direction, the first unit **12** would typically include receiver components (not shown) as will be described below for the second unit **14**, and the second unit **14** would typically include transmitter components (not shown) as described below for the first unit **12**. In accordance with certain embodiments of the present invention, the first unit **12** uses the multiple antenna array **16** for transmitting and receiving communication signals, and the second unit **14** use the multiple antenna array **18** for transmitting and receiving communication signals.

In this particular embodiment for transmissions from the first unit **12** to the second unit **14**, the propagation of the signals over the actual communications channel **10** has a baseband characterization that comprises a matrix of complex-valued (having real and imaginary parts) propagation coefficients between the transmitting antennas **16***a–c* and the receiving antennas **18***a–c*. In this particular embodiment, the first unit **12** and the second unit **14** learn the values of the propagation matrix and determine the appropriate propagation information to establish the virtual sub-channels between the first unit **12** and the second unit **14**. Alternatively, the first unit **12** and/or the second unit **14** do not learn the propagation matrix. Instead, the units **12** and **14** obtain and/or determine some portion of the propagation information, such as the diagonal matrix D, and the first unit **12** obtains and/or determines other propagation information, such as the unitary matrix Φ^+ while the second unit **14** obtains and/or determines the unitary matrix ψ.

To learn the propagation matrix (or derive the appropriate propagation information for certain embodiments) for communications from the first unit **12** to the second unit **14**, the first unit **12** can send training signals to the unit **14** from which the unit **14** determines or estimates the propagation matrix (or propagation information). The unit **14** can then send the propagation matrix (or propagation information) to the unit **12** by using a conventional communication scheme employing one or multiple antennas at the first and second units **12** and **14**. Alternatively, the unit **14** can send the propagation matrix (or propagation information) to the unit **12** using other communication links, such as a phone line. Additionally, the unit **14** can simply send back to the unit **12** the training signals received from the unit **12**, and the unit **12** determines the propagation matrix (or propagation information). The unit **12** then sends the propagation matrix (or propagation information) to the unit **14**.

Likewise, to learn the propagation matrix (or propagation information) for signals propagating from the unit **14** to the unit **12**, the unit **14** sends training signals to the unit **12** from which the unit **12** determines or estimates the propagation matrix (or propagation information). Once again, as discussed for determining the propagation matrix (or propagation information) for transmissions from the unit **12** to the unit **14**, the unit **12** can provide and/or derive different forms or amounts of propagation information to the unit **14** depending on the particular embodiment and using different communication schemes.

Two-way transmissions between the unit **12** and the unit **14** can be accomplished using a variety of schemes, such as time division duplex (TDD) or frequency division duplex (FDD). TDD has the advantage that the channel characteristics for signals propagating from the first unit **12** to the second unit **14** are generally the same as the channel characteristics for signals propagating from the second unit **14** to the second station **12**. In the case of TDD, both units **12** and **14** occupy the same frequency band. As such, the propagation matrix for transmissions from the first unit **12** to the second unit **14** are considered transposes of each other due to the principle of reciprocity as described in D. S. Jones, "Acoustic and Electromagnetic Waves," Oxford University Press, 1989, pp. 63–64 and no further exchange of training signals is required (although updates to the propagation matrix or to propagation information can be performed). In FDD for certain embodiments, the second unit **14** transmits to the first unit **12** the propagation matrix (or propagation information) for transmissions from the first unit **12** to the second unit **14**. Additionally, after the appropriate set-up or training scheme is performed for transmissions from the second unit **14** to the first unit **12**, the first unit **12** can transmit to the second unit **14** the propagation matrix (or propagation information) for transmissions from the second unit **14** to the first unit **12**.

With particular reference to FIG. **3**, since the first and second units **12** and **14** know the propagation matrix (or propagation information which can be derived from the propagation matrix or propagation information which can be used to derive the propagation matrix or other propagation information), the units **12** and **14** can effectively decompose

the complicated multiple-antenna channel **10** into multiple independent virtual sub-channels using the propagation matrix (or respective propagation information). In this particular embodiment, the units **12** and **14** derive the multiple virtual sub-channels from the multiple-antenna channel **10** by using propagation information derived from a singular value decomposition of the propagation matrix. After learning the propagation matrix for transmissions from the first unit **12** to the second unit **14**, the first unit **12** provides the propagation matrix H to a singular value decomposition block **22**. The singular value decomposition **22** performs a singular value decomposition of the propagation matrix H which produces some propagation information, including the diagonal matrix D and two unitary matrices Φ and ψ^+ where the superscript $^+$ denotes a conjugate transpose. The number of nonzero diagonal elements in the matrix D represents the number of parallel independent virtual sub-channels.

For transmissions from the first unit **12** to the second unit **14** in this particular embodiment, the first unit **12** provides the diagonal matrix D to a channel coder/modulator **24** to encode and modulate an incoming information stream from an information source **28**. The channel coder/modulator **24** encodes and modulates the information stream to form a plurality of sub-information streams depending on the nonzero values of the diagonal matrix D to produce the virtual transmitted signal. The virtual transmitted signal can be represented as a vector, the respective components of which are transmitted onto respective virtual sub-channels. In this particular embodiment, a unitary transformation block **30** performs a unitary transformation on the virtual transmitted signal by multiplying the virtual transmitted signal with the conjugate transpose of the unitary matrix Φ. Finally, the first unit **12** uses the results of the unitary transformation **30** to produce the baseband version of the actual transmitted signal.

In this particular embodiment, the second unit **14** also provides the propagation matrix H to a singular value decomposition **32** and uses the results from the singular value decomposition **32** to perform a unitary transformation **34** on the actual received signal. In this particular embodiment, the unitary transformation block **34** multiplies the actual received signal with the unitary matrix ψ to produce a virtual received signal. In this particular embodiment, the multiplications by the unitary matrices at the unitary transformations **30** and **34** tend not to amplify noise and are invertible operations, so information should not be lost. The multiplications by the unitary matrices establishes a link, which can be treated as parallel independent virtual sub-channels, between the virtual transmitted signal at the first unit **12** and the virtual received signal at the second unit **14**. The second unit **14** provides the diagonal matrix D to a channel demodulator/decoder **36** which uses the diagonal matrix D to decode and demodulate the virtual received signal vector to form a single information stream which corresponds to the information stream provided to the channel coder/modulator **24** at the first unit **12**. Additionally, the first unit **12** can use the diagonal matrix D to provide enhanced performance by sending more bits on the stronger virtual sub-channels according to the nonzero values of the diagonal matrix. If the amplitude of a nonzero value of the matrix D is below a certain level, the multiple antenna system can advantageously not use the corresponding virtual sub-channel, thereby a subset of the stronger virtual sub-channels can be used. As such, the multiple antenna system achieves higher data rates in a narrow bandwidth by effectively providing parallel independent virtual sub-channels within the same frequency band and enhanced performance.

FIG. **4** shows a block diagram of a transmitter **40** having the multiple antenna array **16** with three antennas **16***a–c* for use in any unit **12** or **14** according to the principles of the present invention. For ease of discussion, the transmitter **40** will be described as being included in the unit **12** (FIG. **3**) because the multiple-element antenna communication system according to the principles of the present invention has been described in the context of transmissions from the first unit **12** (FIG. **3**) to the second unit **14** (FIG. **3**). The units **12** and **14** (FIG. **3**) in this particular embodiment, however, can both transmit and receive signals according to the principles of the present invention. In doing so, the units **12** and **14** exchange training signals to estimate the propagation matrices (or propagation information which can be the actual training signals in certain embodiments) for communications between them. After learning the propagation matrix H for signals propagating from the first unit **12** (FIG. **3**) to the second unit **14** (FIG. **3**), the transmitter **40** provides the propagation matrix H to the singular value decomposition block **22**.

The singular value decomposition block **22** performs a singular value decomposition of the propagation matrix H (which can be done for any H) to yield $H=\Phi\times D\times\psi^+$, where D is a real-valued, non-negative, diagonal matrix and Φ and ψ^+ are unitary matrices. The number of nonzero diagonal elements is less than or equal to the smaller of the number of antennas **16** at the transmitter **40** and at the second unit **14** (FIG. **3**) and represents the number of parallel independent virtual sub-channels that are available within the same frequency band. Additionally, the sizes of the nonzero diagonal elements indicate the relative signal-to-noise ratios of the virtual sub-channels. In accordance with an aspect of the present invention, the transmitter **40** allocates power to the virtual sub-channels depending on the relative signal-to-noise ratios of the sub-channels as determined by the values of the nonzero diagonal elements of the diagonal matrix D. Thus, the better sub-channels can get more transmitter power and carry more data. In accordance with particular embodiments of this aspect of the present invention, the transmitter **40** has a total power restriction and generally allocates power to the different sub-channels to provide a higher, reliable bit rate. In doing so, the transmitter **40** can allocate power to the different subchannels using a version of a Water Pouring algorithm as disclosed by Robert G. Gallager, "Information Theory and Reliable Communication," John Wiley & Sons, 1968, pp. 343–345.

The transmitter **40** receives information signals from the information source **28**, and the information stream is input into a buffer **43**. According to certain aspects of the present invention, if all of the non-zero diagonal elements of the diagonal matrix D are equal or relatively similar in this particular embodiment, the channel coder/modulator **24** can direct the first bit of a bit stream **42** from the buffer **43** to encoder/modulator **44***a*, the second bit of the bit stream **42** to encoder/modulator **44***b*, and the third bit of the bit stream **42** to encoder/modulator **44***c*. If the amplitude of the nonzero diagonal value of the matrix D corresponding to a first virtual sub-channel is twice as large as the nonzero diagonal values of the matrix D corresponding to the second and third virtual sub-channels in this particular example, the channel coder/modulator **24** could send the first two bits of the bit stream **42** to the encoder/modulator **44***a* and the next two bits of the bit stream **42** being split between the encoder/modulator **44***b* and the encoder/modulator **44***c*. To combat the effects of noise, the channel coder/modulator **24** can use ordinary error-correcting codes of the types typically used for conventional additive Gaussian noise channels as would

be understood by one of ordinary skill in the art in combination with conventional modulation schemes, such as Quadrature Phase Shift Keying (QPSK) modulation.

In this particular embodiment, the channel coder/modulator **24** produces three signal vector components y_{t1}, y_{t2} and y_{t3} of the virtual transmitted signal. The components of the virtual transmitted signal are digital complex values which can be fixed point or floating point depending on the implementation. The three components y_{t1}, y_{t2} and y_{t3} of the virtual transmitted signal from the encoder/modulators **44***a–c* are provided to the unitary transformation block **30**. The unitary transformation block performs a 3×3 matrix multiplication on the components y_{t1}, y_{t2} and y_{t3} using a unitary matrix Φ^+. The result of the unitary transformation **30** is an actual transmitted signal having components s_{t1}, s_{t2} and s_{t3} which are sent to respective transmit circuits **48***a–c* and converted as necessary to the radio frequency (RF) domain. In this particular embodiment, the transmit circuitry **48***a–c* modulates the components s_{t1}, s_{t2} and s_{t3} of the actual transmitted signal onto the same carrier and transmits each component s_{t1}, s_{t2} and s_{t3} of the actual transmitted signal over a respective antenna **16***a–c*.

FIG. **5** shows a general diagram for an embodiment of the transmit circuitry **48**. In this particular embodiment, each components s_{t1}, s_{t2} and s_{t3} of the virtual transmitted signal vector is split into real and imaginary parts by block **50**. The real and imaginary parts of each component s_{t1}, s_{t2} and s_{t3} are input to digital-to-analog converters **51**. Multipliers **52** multiply the analog real part of each component s_{t1}, s_{t2} and s_{t3} by $\cos\omega_c t$ with ω_c being the carrier frequency (radians/second) and multiply the analog imaginary part of each component s_{t1}, s_{t2} and s_{t3} by $\sin\omega_c t$. Afterward, respective real and imaginary parts of each component s_{t1}, s_{t2} and s_{t3} are added together by summers **53** to produce an RF signal for each component s_{t1}, s_{t2} and s_{t3} on the same carrier frequency ω_c. Each RF signal is then amplified by respective RF power amplifiers **54** and transmitted over respective antennas **16***a–c* in this particular embodiment. Alternative embodiments for the transmit circuitry **48** are possible.

FIG. **6** shows a block diagram of a receiver **60** having the multiple antenna array **18** with three antennas **18***a–c* for use in any unit **12** or **14** (FIG. **3**) according to the principles of the present invention. For ease of discussion, the receiver **60** will be described as being included in the second unit **14** (FIG. **3**) because the multiple-element antenna communication system according to the principles of the present invention has been described in the context of transmissions from the first unit **12** (FIG. **3**) to the second unit **14** (FIG. **3**). As mentioned above, the units **12** and **14** (FIG. **3**) in this particular embodiment can both transmit and receive signals according to the principles of the present invention.

As described above for this particular embodiment, the receiver **60** receives training signals from the transmitter **40** (FIG. **4**) and estimates the propagation matrix H for communications from the transmitter **40** (FIG. **4**) to the receiver **60**. After learning the propagation matrix H for signals propagating from the transmitter **40** (FIG. **4**) to the receiver **60**, the receiver **60** provides the propagation matrix H to the singular value decomposition block **32**. As in the transmitter **40** (FIG. **4**), the singular value decomposition block **32** performs a singular value decomposition of the propagation matrix H to yield $H=\Phi\times D\times\psi^+$, where D is a real-valued, non-negative, diagonal matrix and Φ and ψ^+ are 3×3 complex unitary matrices with the superscript "+" denoting the "conjugate transpose."

The receiver **60** receives the signals propagating from the transmitter **40** (FIG. 4) through the multiple antenna array **18**, and receiver circuits **62***a–c* process the signals received from the respective antennas **18***a–c* down to baseband. In this particular embodiment, the actual received signal vector $\underline{x}_t$ is digitized before being provided to the unitary transformation block **34**. The unitary transformation block **34** uses its propagation information derived from the propagation matrix H to multiply the actual received signal $\underline{x}_t$ by the 3×3 unitary matrix ψ to obtain a virtual received signal $\underline{r}_t=[r_{t1}, r_{t2}, r_{t3}]=\underline{x}_t\times\psi$. The channel demodulator/decoder **36** receives the virtual received signal $\underline{r}_t$ from the unitary transformation block **34** with respective parallel components r_{t1}, r_{t2} and r_{t3}. The components r_{t1}, r_{t2} and r_{t3} are provided to respective parallel demodulators/decoders **68***a–c*. The parallel demodulators/decoders **68***a–c* demodulate and decode the virtual received signal $\underline{r}_t$ according to the modulation and coding scheme used by the transmitter **40** (FIG. **4**).

Using the nonzero diagonal values of the diagonal matrix D from the SVD block **32** and reflecting the use of the matrix D in the transmitter **40** (FIG. **4**), the channel demodulator/decoder **36** constructs a single stream **70** of information bits from the parallel components r_{t1}, r_{t2} and r_{t3} of the virtual received signal. As such, the diagonal matrix D used to construct the information stream **70** from the parallel streams **64***a–c* in the receiver **60** is the same matrix D used to separate the single stream **42** (FIG. **4**) into parallel streams **46***a–c* (FIG. **4**) in the transmitter **40** (FIG. **4**). The information stream **70** is then output to its destination **38** before which the information stream **70** can pass through additional processing or circuitry **72** which can include a data buffer.

FIG. **7** shows a general diagram for an embodiment of the receive circuitry **62**. In this particular embodiment, the transmitted RF signals from the transmit circuitry **48** (FIG. **5**) are received at the antennas **18***a–c*. The RF signals received at each antenna **18***a–c* are amplified by respective pre-amplifiers **80**. Multipliers **82** multiply the respective RF signals by $\cos\omega_c t$ and $\sin\omega_c t$ with ω_c being the carrier frequency to produce analog versions of the real and imaginary parts of the components of the actual received signal which were modulated onto carriers of the same frequency ω_c. The analog versions of the real and imaginary parts for each component of the actual received signal are low pass filtered by filters **83** and provided to analog-to-digital converters **84** to produce digital versions of the real and imaginary parts of each component. Combiner **86** combine the real and imaginary parts of each component to produce the components x_{t1}, x_{t2} and x_{t3} of the actual received signal as complex digital values. In this particular embodiment as described above, the components x_{t1}, x_{t2} and x_{t3} of the actual received signal are provided to the unitary transformation block **34** (FIG. **6**). The receive circuitry **62** is shown as a homodyne receiver, but alternative embodiments for the receive circuitry **62** are possible.

In accordance with certain embodiments of the present invention, the multiple antenna system updates the propagation matrix or information periodically or continuously as operating conditions change. As such, if an antenna **16** (FIG. **4**) or **18** becomes unavailable, the multiple antenna system can update the propagation matrix(ices) or information at the two units **12** and **14**, thereby maintaining the communications between the units **12** and **14** and still providing a high bit rate. Additionally, depending on the conditions of the virtual sub-channels (which can be measured by the nonzero diagonal values of the diagonal matrix D), the bit rate between the two units **12** and **14** using the multiple antenna system is scaleable. Accordingly, if the signal-to-noise ratio, virtual sub-channel gain or other measurement value for a virtual sub-channel drops below a certain threshold level

and/or a relative level compared to the other sub-channels, the multiple antenna system can either reduce the number of bits transmitted over the virtual sub-channel or drop the virtual sub-channel, thereby reducing the bit rate, until updates of the propagation matrix or information show that the virtual sub-channel has risen above the threshold level and/or the relative level. Furthermore, in similar fashion, the multiple antenna system according to certain principles of the present invention can allocate power to the different virtual sub-channels based on the conditions of the different virtual sub-channels as determined by the nonzero values of the diagonal matrix D or other measurement values corresponding to the virtual sub-channels.

The multiple antenna system enables an increase in bit rate without an increase in power or bandwidth as compared to single antenna systems. In certain embodiments, the propagation of the signals is modeled as flat fading (no frequency dependence to fading). Additionally, in the event of receiver noise and external interference, the multiple antenna system can use a covariance matrix or the like which characterizes the receiver noise and/or the external interference to effectively alter the propagation information or to include the covariance matrix or the like as part of the propagation information. If the elements of the propagation matrix H are statistically independent, identically distributed, with Rayleigh-distributed magnitude and uniformly distributed phase, the capacity of the channel grows linearly with the smaller of the number of transmitter and receiver antennas. Theoretically, there is no limit to the number of antennas that can be utilized in the multiple antenna system, thereby providing the potential for enormous capacities in a narrow bandwidth. For example, if 170 antennas are used at the transmitter and the receiver in a 30 kHz bandwidth with signal-to-noise ratios of 20 dB, 20 Mb/s of High Definition TV (HDTV) may be sent over the multiple antenna channel after being decomposed to 170 virtual sub-channels according to the principles of the present invention.

The multiple-element antenna system according to the principles of the present invention can achieve such high bit rates in the narrow bandwidth and in a simple manner by using multiple antennas at both the transmitter and receiver to decompose the multiple antenna channel into multiple independent virtual sub-channels. A pair of units using the multiple antenna system use a distinct band of frequencies for communicating with each other over the multiple-antenna channel. Thus, different pairs of units can be frequency division multiplexed. A pair of units can take a variety of forms, such as units which can handle voice and/or data in a wireless local area network. Additionally, the multiple antenna system can provide security in communications between a pair of units because the propagation information characterizing the actual communications channel between the pair of units is unique to that pair. As such, even if a third party obtained the propagation information for the pair, it would be difficult for the third party to intercept the communications between the pair because the propagation information for an actual communications channels between the third party and each of the pair of units would be different.

In addition to the embodiments described above, alternative configurations of the multiple antenna system according to the principles of the present invention are possible which omit and/or add components and/or use variations or portions of the described multiple antenna system. For example, the multiple antenna system has been described with three antennas at both the first unit and the second unit to provide three virtual sub-channels, but different numbers of antennas can be employed at the first unit and the second unit. Additionally, the multiple antenna system has been described as being employed in terms of communications over a multiple antenna channel from the first unit to the second unit. The multiple antenna system, however, includes units which transmit and/or receive according to the principles of the present invention. In certain embodiments, the units use the same multiple antennas for both transmission and reception. Alternative embodiments of units employing the multiple antenna system are possible, however, which use a subset of the multiple antennas for transmission and/or reception depending on the number of antennas at the unit on the other end of the multiple antenna channel or upon the values of the diagonal matrix in the reverse direction.

The above-described multiple antenna system has been described as comprising several components or blocks, but it should be understood that the multiple antenna system and portions thereof can be implemented in application specific integrated circuits, software-driven processing circuitry, or other arrangements of discrete components as would be understood by one of ordinary skill in the art with the benefit of this disclosure. What has been described is merely illustrative of the application of the principles of the present invention. Those skilled in the art will readily recognize that these and various other modifications, arrangements and methods can be made to the present invention without strictly following the exemplary applications illustrated and described herein and without departing from the spirit and scope of the present invention.

We claim:

1. A method for transmitting communication signals, said method comprising:

sending by a first unit a virtual transmitted signal over at least a subset of virtual sub-channels of an actual communications channel using at least a portion of propagation information characterizing said actual communications channel between M transmitter antennas of said first unit and N receiver antennas of a second unit, where M or N>1.

2. The method of claim **1** further comprising:

transmitting by said first unit training signals to said second unit; and

obtaining said at least a portion of said propagation information by said first unit from said second unit.

3. The method of claim **1** further comprising:

transmitting by said first unit training signals to said second unit; and

obtaining by said first unit training signals as received at said second unit for determining said at least a portion of said propagation information.

4. The method of claim 1 wherein said sending further including:

transforming said virtual transmitted signal into an actual transmitted signal using said at least a portion of said propagation information; and

transmitting said actual transmitted signal over said M transmitter antennas onto said actual communications channel.

5. The method of claim **4** wherein said transmitting further including:

transmitting each component of said actual transmitted signal using a respective one of said M transmitter antennas.

6. The method of claim **4** further including:

producing a unitary matrix from said at least a portion of said propagation information.

7. The method of claim 6 wherein said producing further includes:

performing a singular value decomposition of a propagation matrix H which characterizes said actual communications channel.

8. The method of claim 6 wherein said producing further including:

using eigenvectors of a matrix derived from a propagation matrix H which characterizes said actual communications channel.

9. The method of claim 6 further including:

using said unitary matrix to perform a unitary transformation of said virtual transmitted signal into said actual transmitted signal; and

transmitting each component of said actual transmitted signal onto carriers having the same frequency.

10. The method of claim 1 further including:

receiving by the first unit an information stream from an information source; and

separating said information stream into a plurality of signal streams using at least a portion of said propagation information to form said virtual transmitted signal.

11. The method of claim 10 further including:

producing a diagonal matrix D from at least said portion of said propagation information with the nonzero diagonal values corresponding to said virtual sub-channels of said actual communications channel.

12. The method of claim 11 wherein said producing further includes:

performing a singular value decomposition of a propagation matrix H which characterizes said actual communications channel.

13. The method of claim 12 wherein said producing further including:

using squareroots of eigenvalues of a matrix derived from a propagation matrix H which characterizes said actual communications channel.

14. The method of claim 10 wherein said separating further including:

using values of a diagonal matrix D from at least said portion of said propagation information to separate said information stream into a plurality of signal streams according to said values of said diagonal matrix D.

15. The method of claim 14 further including:

using said values of said diagonal matrix D to allocate power for transmitting each component of said virtual transmitted signal.

16. A method of receiving communication signals, said method comprising:

retrieving by said second unit a virtual received signal from at least a subset of virtual sub-channels of an actual communications channel using at least a portion of propagation information characterizing said actual communications channel between M transmitter antennas of a first unit and N receiver antennas of said second unit, where M or N>1 and said first unit includes at least a portion of said propagation information.

17. The method of claim 16 further comprising:

receiving by said second unit training signals from said first unit; and

determining said at least a portion of said propagation information by said second unit using said training signals received by said second unit.

18. The method of claim 16 further comprising:

receiving by said second unit training signals from said first unit; and

sending training signals as received by said second unit to said first unit;

determining by said first unit said at least a portion of said propagation information using said training signals as received by said second unit; and

sending said at least a portion of said propagation information by said first unit to said second unit.

19. The method of claim 16 wherein said retrieving further including:

receiving an actual received signal from said actual communications channel on said N receiver antennas; and

transforming said actual received signal into said virtual received signal using said at least a portion of said propagation information.

20. The method of claim 19 wherein said retrieving further including:

receiving each component of said actual received signal using a respective one of said N receiver antennas.

21. The method of claim 19 further including:

producing a unitary matrix from said at least a portion of said propagation information.

22. The method of claim 21 wherein said producing further includes:

performing a singular value decomposition of a propagation matrix H which characterizes said actual communications channel.

23. The method of claim 21 wherein said producing further including:

using eigenvectors of a matrix derived from a propagation matrix H which characterizes said actual communications channel.

24. The method of claim 21 further including:

receiving each component of said actual received signal from carriers having the same frequency; and

using said unitary matrix to perform a unitary transformation of said actual received signal into said virtual received signal.

25. The method of claim 16 further including:

combining components of said virtual received signal to provide an information stream using at least a portion of said propagation information.

26. The method of claim 25 wherein said combining further including:

producing a diagonal matrix D from at least said portion of said propagation information.

27. The method of claim 26 wherein said producing further includes:

performing a singular value decomposition of a propagation matrix H which characterizes said actual communications channel.

28. The method of claim 26 wherein said producing further including:

using squareroots of eigenvalues of a matrix derived from a propagation matrix H which characterizes said actual communications channel.

29. The method of claim 25 wherein said combining further including:

using values of a diagonal matrix D from at least said portion of said propagation information to combine said virtual received signal into an information stream according to said values of said diagonal matrix D.

30. A method of communicating over an actual communications channel, said method comprising:

creating virtual sub-channels from said actual communications channel between M transmitter antennas of a first unit and N receiver antennas of a second unit, where M or N>1, by using at said first unit and said second unit propagation information characterizing said actual communications channel;

sending by said first unit a virtual transmitted signal over at least a subset of said virtual sub-channels using at least a portion of said propagation information; and

retrieving by said second unit a virtual received signal from said at least a subset of said virtual sub-channels using at least another portion of said propagation information.

31. A transmitter for transmitting information signals comprising:

a plurality of antennas;

processing circuitry configured to obtain propagation information for an actual communications channel between said plurality of antennas and a plurality of receiver antennas, said processing circuitry comprising a channel coder configured to receive an information signal stream and at least a portion of said propagation information, said channel coder further configured to separate said information signal stream using said at least a portion of said propagation information to form a virtual transmitted signal of components corresponding to sub-channels of said communications channel, said processing circuitry further configured to perform a transformation on said virtual transmitted signal using at least another portion of said propagation information to form an actual transmitted signal; and

transmit circuitry coupled to said plurality of antennas and configured to transmit each component of said actual transmitted signal through a respective one of said plurality of antennas on carriers of the same frequency.

32. A receiver for receiving information signals comprising:

a plurality of antennas;

receive circuitry coupled to said plurality of antennas and configured to receive each component of an actual received signal through a respective one of said plurality of antennas; and

processing circuitry configured to obtain propagation information for a communications channel between said plurality of antennas and a plurality of transmitter antennas at a transmitter having at least a portion of said propagation information, said processing circuitry further configured to perform a transformation on said actual received signal using at least a portion of said propagation information to form a virtual received signal, said processing circuitry comprising a channel decoder configured to receive said virtual received signal and at least another portion of said propagation information, said channel decoder further configured to combine said virtual received signal into an information stream using said at least another portion of said propagation information.

33. A method for transmitting communication signals on an actual communications channel between M transmitter antennas of a first unit and N receiver antennas of a second unit, where M or N>1, said method comprising:

producing by said first unit using at least a portion of propagation information characterizing said actual communications channel an actual transmitted signal of components for transmission over said actual communications channel to said second unit for transformation by said second unit using at least a portion of said propagation information.

34. A method of receiving communication signals from an actual communications channel between M transmitter antennas of a first unit and N receiver antennas of a second unit, where M or N>1, said method comprising:

transforming by said second unit using at least a portion of propagation information characterizing said actual communications channel a plurality of components of an actual received signal received from said actual communications channel from said first unit after being produced at said first unit using at least a portion of said propagation information.

35. A method for transmitting communication signals, said method comprising:

creating virtual sub-channels from an actual communications channel between M transmitter antennas of a first unit and N receiver antennas of a second unit, where M or N>1, by using at said first unit and said second unit at least portions of said propagation information characterizing said actual communications channel; and

sending by said first unit a virtual transmitted signal over at least a subset of said virtual sub-channels using at least a portion of said propagation information.

36. A method of receiving communication signals, said method comprising:

creating virtual sub-channels from an actual communications channel between M transmitter antennas of a first unit and N receiver antennas of a second unit, where M or N>1, by using at said first unit and said second unit at least portions of said propagation information characterizing said actual communications channel; and

retrieving by said second unit a virtual received signal from at least a subset of said virtual sub-channels using at least a portion of said propagation information.

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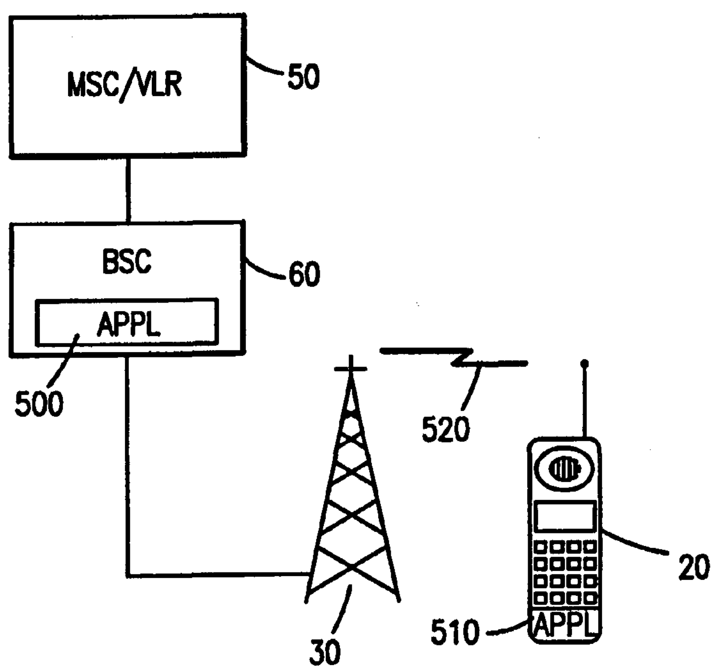
(57) Abstract

A broadcast message indicating the utilization level associated with Stand-alone Dedicated Control Channels (SDCCH) within a mobile telecommunications network serving a particular geographic area is transmitted over a broadcast channel. A plurality of mobile stations located within that particular geographic area monitoring the broadcast channel then receives the transmitted message. Thereafter, a mobile service request with a lower priority level than the indicated status level is queued by the associated mobile station and delayed until the channel utilization level rescinds to a network acceptable level.



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OPTIMAL USE OF LOGICAL CHANNELS WITHIN A MOBILE TELECOMMUNICATIONS NETWORK

BACKGROUND OF THE INVENTION

Technical Field of the Invention

The present invention relates to a telecommunications network and, in particular, to the efficient management of channel resources within a digital mobile communications network.

Description of Related Art

The general name of the connection between a particular mobile station traveling within a particular cell area and the base transceiver station (BTS) providing radio coverage for that particular cell area is the "radio interface" or "air interface". Historically, the communications of information across the air interface between a base transceiver station (BTS) and a mobile station has employed, so-called, analog modulation techniques. For example, Frequency Division Multiple Access (FDMA) technology has been widely utilized to assign each mobile station to one of a plurality of the frequency channels associated with the current cell area to communicate with the serving BTS. More recently, however, digital modulation techniques have been used in order to enhance the spectrum efficiency with which the bandwidth allotted to mobile communications is used. As an illustration, the two techniques of time division multiple access (TDMA) and code division multiple access (CDMA) have been utilized to allow communications to proceed between a BTS and a plurality of different mobile stations on a relatively limited amount of radio frequency bandwidth. The Global System for Mobile (GSM) communications system, for example, utilizes the TDMA concept with the allocation of one TDMA frame per carrier frequency channel to communicate between a mobile station and a BTS. Each frame is further subdivided into eight

time-slots (TS). Each time-slot of a TDMA frame on a single frequency channel is referred to as a physical channel. Accordingly, there are eight physical channels per carrier in the GSM system. Each physical channel of the GSM system can be compared with one single channel in an FDMA-system, where every user is connected to the system via one of the associated frequencies.

The implementation of TDMA technology requires that a great quantity and variety of information must be transmitted between the serving BTS and the mobile station over the limited physical channels. For example, control data, service request data, actual traffic data, supplementary data, etc., have to be communicated over the physical channels. As a result, in order to distinguish one type of data from another, different logical channels have been named and mapped (assigned) on to the available physical channels. For example, actual speech is sent on the logical channel named "traffic channel (TCH)" occupying one or more physical channels. Paging of a called party mobile station is performed over the logical "paging channel (PCH)" while synchronization of a mobile station with a serving BTS is performed over the logical "synchronization channel (SCH)" which occupies one of the physical channels. Accordingly, depending on the type of information being transmitted, different logical channels are utilized. Needless to say, if more physical channels are assigned to a particular logical channel, a lesser number of physical channels are available for the rest of the logical channels.

Because of the limited physical channel resources, mobile service providers are often faced with channel resource management and dimensioning problems. One such problem includes effectively managing Stand-alone Dedicated Control Channels (SDCCH) within a serving mobile network. Because of the fact that congestion in SDCCH logical channels results in lost calls and unsuccessful call setups, the efficient management of SDCCH logical

channels is critical for providing reliable mobile service to mobile stations traveling within the serving coverage area.

SDCCH logical channels are not only utilized for setting up call connections, but also for performing location updates for traveling mobile stations and for communicating packet messages containing text or graphic data between the serving mobile network and associated mobile stations. Conventionally, all of the above mentioned functionalities are provided the same priority and allowed equal access to available SDCCH channel resources. As a result, all of the available SDCCH channel resources could be occupied by Short Message Service (SMS) or Unstructured Supplementary Service Data (USSD) messages transporting text messages and could, as a result, prevent speech connections from being established between mobile stations and a serving mobile network. However, even though the support of all of the above mentioned functionalities is important for providing reliable and comprehensive mobile service to associated mobile stations, establishing call connections is by far the most important role performed by the mobile network.

Accordingly, there is a need for a mechanism to prioritize mobile services within a serving mobile network to better utilize available SDCCH logical channels.

SUMMARY OF THE INVENTION

The present invention discloses a method and apparatus for optimizing the utilization of Stand-alone Dedicated Control Channels (SDCCH) within a mobile telecommunications network for high priority mobile services. The level of utilization of SDCCH logical channels is maintained within a particular base station controller (BSC) serving a particular geographic area. After determining that such a level has exceeded a threshold value imposed by the serving telecommunications network, the BSC transmits data over a Broadcast Control

Channel informing associated mobile stations traveling within its coverage area of such a determination. Thereafter, mobile stations wanting to request low priority mobile services are instructed to delay requesting one of the available SDCCH logical channels until the utilization level falls below the imposed threshold level. As a result, available SDCCH logical channels remain available for high priority mobile services within a congested mobile telecommunications network.

BRIEF DESCRIPTION OF THE DRAWINGS

A more complete understanding of the method and apparatus of the present invention may be had by reference to the following detailed description when taken in conjunction with the accompanying drawings wherein:

FIGURE 1 is a block diagram of a mobile telecommunications network illustrating a mobile station communicating with a serving base transceiver station (BTS);

FIGURE 2 is a block diagram of physical channels allocated in accordance with the Time Division Multiple Access (TDMA) technology;

FIGURE 3 is a block diagram of different logical channels within a TDMA physical frame in accordance with the Global System for Mobile (GSM) standard;

FIGURE 4 is a signal sequence diagram illustrating the communication of different messages for originating an outgoing call connection;

FIGURE 5 is a block diagram of a serving mobile network transmitting Stand-alone Dedicated Control Channel (SDCCH) status information to a mobile station in accordance with the teachings of the present invention; and

FIGURE 6 is a flow-chart illustrating the steps performed by a mobile station to request mobile service towards a serving mobile network.

DETAILED DESCRIPTION OF THE DRAWINGS

FIG. 1 is a block diagram of a Public Land Mobile Network (PLMN) 10 illustrating a mobile station 20 communicating with a serving base transceiver station (BTS) 30. A geographic area associated with a particular Public Land Mobile Network (PLMN) 10 is partitioned into a number of smaller areas. Whenever a mobile station 20 travels into one of those smaller areas known as a "location area", the mobile station performs a location update with the serving PLMN. Such a location update informs the associated mobile switching center / visitor location register (MSC/VLR) 50 of the mobile station's presence. In case the mobile station 20 is an unregistered subscriber, a home location register (HLR) 60 associated with the newly registering mobile station 20 is identified and necessary communication is facilitated between the serving MSC/VLR 50 and the HLR 60 to authenticate the new mobile station 20. Requisite subscriber information related to the newly registering mobile station 20 is further requested and retrieved from the associated HLR 60 and stored at the serving MSC/VLR 50. Thereafter, the mobile station 20 is allowed to access mobile service within the serving MSC/VLR 50 coverage area.

Whenever an incoming call connection is requested towards the mobile station 20, a call setup signal, such as an Integrated Service Digital Network User Part (ISUP) based Initial Address Message (IAM) is received by a gateway mobile switching center (GMSC) 80 associated with the HLR 60. After performing HLR interrogation to ascertain the current location of the mobile station 20, the received incoming call setup signal is rerouted by the GMSC 80 to the MSC/VLR 50 currently serving the mobile station 20. The MSC/VLR 50 then determines the current location area of the mobile station 20 and instructs the appropriate base station controller (BSC) 40 to page the mobile station 20. The BTS 30 then pages the mobile

station to alert the mobile station of an incoming call. As soon as the mobile station detects the paging message, the mobile station 20 sends a request for a signaling channel to the BSC 40. After allocating an idle signaling channel to the mobile station 20, the BSC 40 sends a message instructing the mobile station 20 to switch to that particular signaling channel. After communicating necessary control and service related data over the newly allocated signaling channel, a speech channel is subsequently seized and a call connection between the mobile station 20 and the BSC 40 is established via the BTS 30.

Initially, the communication interface 90 between the serving BTS 30 and the mobile station 20 employed so called analog modulation techniques. However, with the recent developments of digital communication technology, digital modulation techniques are being used to enhance the efficiency and capacity of data communications within a mobile telecommunications network. As an illustration, the techniques of time division multiple access (TDMA) or code division multiple access (CDMA) are being used to allow multiple communications to proceed on a relatively limited amount of radio frequencies. Global System for Mobile (GSM) based telecommunications networks, for example, utilize the TMDA technology with one TDMA frame per carrier frequency to communicate between a mobile station and a BTS.

Reference is now made to FIG. 2 depicting a diagrammatic representation of time-frame structures within the GSM standard. The longest recurrent time period of the structure is called a hyperframe 100 and has the duration of 3 hours 28 minutes 53 seconds 760 ms. One hyperframe 100 is divided into 2048 superframes 105, each having a duration of 6.12 seconds. The superframe 105 is itself sub-divided into a number of multiframes. Two types of multiframes exist in the GSM standard. First, there is a fifty-one (51) frame multiframe 110 with a

duration of 120 ms, comprising twenty-six (26) TDMA frames 130. Next, there is a twenty-six (26) frame multiframe 120 with a duration 235.4 ms, comprising fifty-one (51) TDMA frames 140. Lastly, each TDMA frame within a multiframe has eight time slots 150. Each of these eight physical time slots is equivalent to one Frequency Division Multiple Access (TMDA) channel serving a single mobile station.

A great quantity and variety of information must be transferred between the BTS and the mobile station. For example, paging to inform the mobile station of an incoming call has to be performed over one of the time slots. A request for mobile service further needs to be communicated over one of the time slots. Furthermore, the actual voice data must be communicated over the available time slots. Therefore, in order to distinguish one type of information over another, different logical channels have been introduced and assigned to each of the eight physical time slots.

Reference is now made to FIG. 3 illustrating different logical channels within the GSM standard which can be separated into two broad categories: traffic channels (TCH) 160 and signaling channels 170. Traffic channels (TCH) 169 are utilized by the serving BSC to communicate call data (e.g., voice data) with a particular mobile station traveling within its coverage area. On the other hand, signaling channels 170 are utilized by the serving BSC and BTS to communicate other control data necessary to implement the communication of call data with the mobile station.

Signaling channels are further subdivided into three categories: broadcast control channels 270, common control channels 280, and dedicated control channels 280. Each of the above three categories are then still further sub-divided into a number of logical channels for transporting different types of information between the serving BTS and the mobile station.

Broadcast control channels 270 are mainly utilized for communicating information from the serving BTS to a particular mobile station traveling within its coverage area (down-link) and include the Frequency Correction Channel (FCCH) 180, Synchronization Channel (SCH) 190, and Broadcast Control Channel (BCCH) 200. The Frequency Correction Channel (FCCH) 180 carries information for frequency correction of the mobile station. The Synchronization Channel (SCH) 190 carries information for frame synchronization of the mobile station and identification of the BTS. Lastly, the Broadcast Control Channel (BCCH) 200 is used to broadcast general system information about the cell to all mobile stations located within its location area. For example, the broadcast system information includes data about the network that the mobile station needs to be able to communicate with the network in an appropriate manner. Such information includes cell description, location area identity, neighbor cell description, etc.

Common control channels 280 include the Paging Channel (PCH) 210, Random Access Channel (RACH) 220, and Access Grant Channel (AGCH) 230. The Paging Channel (PCH) 210 is used on the downlink to page a mobile station. For example, when an incoming call setup request is received by the serving MSC/VLR, the appropriate BSC currently serving the mobile station is instructed to page the specified mobile station over a PCH. The Random Access Channel (RACH) 220, on the other hand, is used by the mobile station to request allocation of a Stand-alone Dedicated Control Channel (SDCCH) 240 to the BSC. For example, upon detecting the paging message informing the mobile station of an incoming call, the called party mobile station requests a SDCCH from the serving BSC over a RACH. After allocating an idle SDCCH, the BSC utilizes an Access Grant Channel (AGCH) 230 to communicate the identity of the allocated SDCCH to the requesting mobile station.

Dedicated control channels 290 include the Stand-alone Dedicated Control Channel (SDCCH) 240, Slow Associated Control Channel (SACCH) 250, and the Fast Associated Control Channel (FACCH) 260. The Stand-alone Dedicated Control Channel (SDCCH) 240 is used for signaling with a dedicated mobile station. Accordingly, the SDCCH 240 is the channel used for performing location update procedures whenever a mobile station enters a new location area. The SDCCH is also utilized to initiate a call setup and to seize a TCH. Furthermore, SDCCH logical channels are utilized by the serving mobile network to communicate Unstructured data, such as Short Message Service (SMS) or Unstructured Supplementary Service Data (USSD) messages with associated mobile stations. The Slow Associated Control Channel (SACCH) 250 is associated with a TCH 160 or an SDCCH 240. The SACCH 250 is a continuous data channel carrying continuous control information, such as measurement reports, timing advance and power order, between the serving BSC and the mobile station. Lastly, the Fast Associated Control Channel (FACCH) 260 is associated with a particular TCH to work in burst stealing mode to replace speech or data traffic with other necessary signaling.

As illustrated above, with nine different types of logical signaling channels and one logical traffic channel occupying the limited physical channels, the eight time slots within a TDMA frame need to be managed efficiently and effectively to provide reliable mobile service to mobile stations traveling within a particular BSC coverage area. Since logical channel assignments to physical channels can not be changed dynamically as demands for each logical channel changes within a serving network, determining the appropriate number of physical time slots to be assigned to each of the logical channels is crucial. Especially since congestion in the two of the most frequently utilized logical channels (SDCCH and TCH) results in failed call connection and lost calls.

Therefore, even after allocating an appropriate number of physical channels to each logical channel, efficient management of channel resources is further necessary to maximize the potential utilization of available logical channels.

FIG. 4 is a signal sequence diagram illustrating a normal call setup procedure in accordance with the GSM standard. Whenever a mobile station 20 requests mobile service towards the serving mobile switching center (MSC) 50, either for originating an outgoing call connection or for receiving an incoming call connection, the mobile station 20 transmits a channel request message over a Random Access Channel (RACH) towards the connected BSC 40. After communicating with the associated MSC 50, an available SDCCH channel is allocated. The serving BSC 40 then transmits an Immediate Assignment Command message 310 to the serving BTS 30 (not shown in FIG. 4) to assign the allocated SDCCH to the requesting mobile station 20. The Immediate Assignment message 310 is further transmitted to the requesting mobile station 20 to instruct the mobile station 20 to switch to the assigned SDCCH. Utilizing the assigned SDCCH logical channel, the mobile station 20 requests mobile service from the serving MSC 50 by transmitting a Ciphering Mode Service Request(CM Serv. Req.) message 320. The serving BSC 40 then sets up an Signaling Connection Control Part (SCCP) connection with the MSC 50 by transmitting a SCCP - Connection Request (CR) message 330. The received CM-SERV. REQ. message 320 may further be "piggy-backed" to the transmitted SCCP-CR message 330. The serving MSC 50 is then aware of the mobile station's request for mobile service. As a result, the MSC 50 attempts to authenticate the mobile station 20 by transmitting an Authentication Request message 350 to the mobile station 20 transparently through the connected BSC 40 and over the assigned SDCCH logical channel. The mobile station 20, in response, attempts to comply with the authentication process by returning an Authentication

Response message 360 to the serving MSC 50. If the authentication procedure is successfully performed, the ciphering mode setting procedure may be initiated by the serving MSC 50. Accordingly, the serving MSC 50 sends a Ciphering Mode Command (not shown in FIG. 4) to BSC 40. The BSC 40, in turn, transmits a Ciphering Mode Command message 370 including a cipher key to the mobile station 20 over the assigned SDCCH logical channel. The provided cipher key is then later utilized by the serving BTS and the mobile station to cipher and decipher digital data transmitted over the radio interface. After storing the received cipher key, the mobile station 20 returns a Ciphering Mode Complete message 380 to the serving BSC 40 over the assigned SDCCH logical channel. The received Ciphering Mode Complete message 390 is then transmitted to the serving MSC 50. Furthermore, in order to guarantee the mobile subscriber's confidentiality, instead of identifying the mobile station with its permanently assigned International Mobile Subscriber Identity (IMSI) number, a Temporary Mobile Subscriber Identity (TMSI) number is further assigned by the serving MSC 50. The assigned TMSI number is then included in a TMSI Reallocation Command message 400 and transmitted to the mobile station 20 over the assigned SDCCH logical channel. The mobile station 20 then confirms the receipt of the assigned TMSI number by transmitting a TMSI Reallocation Complete message 410 over the assigned SDCCH logical channel back to the serving MSC 50. The mobile station 20 is now prepared to originate an outgoing call connection and transmits a Call Setup message 420 towards the serving MSC 50. The transmitted Setup message 420, for example, includes the directory number associated with the intended called party subscriber. The serving MSC 50 then acknowledges the call setup signal by transmitting a Call Confirmation message 430 back to the requesting mobile station 20 over the assigned SDCCH logical channel. The BSC 40 then selects an idle traffic channel (TCH) and

instructs the mobile station 20 to tune to the newly allocated TCH logical channel by transmitting an Assignment Command message 440 over the SDCCH logical channel. As an acknowledgment, the mobile station 20 returns an Assignment Complete message 450 back to the serving MSC 50 indicating that the traffic channel is up and running. The BSC then releases the no-longer-needed SDCCH logical channel. An Alert message 460 is then transmitted from the serving BSC 40 to the mobile station 20 informing the mobile station that a ringing tone has been generated by the serving MSC 50. Thereafter, a Connect signal 470 is transmitted from the mobile station 20 to the serving MSC 50. The serving MSC 50 then acknowledges the connect signal by returning a Connect Acknowledgment message 480 to the requesting mobile station 20. Thereinafter, speech connection is established over the newly allocated TCH logical channel allowing the mobile station 20 to communicate data (e.g., voice) with its called party subscriber.

It is to be understood that the call originating procedure illustrated above is for exemplary purposes only and that call termination procedures for an incoming call setup are also applicable in the sense that the SDCCH is also needed and extensively utilized before a call can be terminated towards its called party mobile station.

As illustrated above, until a TCH logical channel is seized enabling a mobile subscriber to communicate with another telecommunications terminal, a great variety and quantity of information needs to be exchanged between the serving mobile network and the requesting mobile station over a SDCCH logical channel. Such information includes authentication data, channel assignment data, service request data, TMSI number data, and call setup information data. Accordingly, without an available SDCCH channel, even if a TCH logical channel is available, a mobile station is not able to access mobile service. Such mobile service includes not only call setup requests as explained

above but a number of other functionalities. Each time a mobile station travels into a new location area being served by a new BSC, the traveling mobile station needs to perform a location update with its associated home location register (HLR) over a SDCCH logical channel. Periodically, the mobile station then has to inform the serving BSC and MSC that the mobile station is still within the service area by performing International Mobile Subscriber Identity (IMSI) Attach procedures over SDCCH logical channels. Terminating call connections, in a manner similar to as described above, also need SDCCH logical channels to alert mobile stations and to establish terminating call connections. Furthermore, unstructured data, such as Short Message Service (SMS) and Unstructured Supplementary Service Data (USSD) messages are also communicated over SDCCH logical channels. With all of the above described mobile services attempting to seize and utilize a limited number of SDCCH logical channels within a particular mobile telecommunications network, providing efficient and effective SDCCH logical channel resource management becomes crucial for providing reliable and efficient mobile service to associated mobile stations.

Reference is now made to FIG. 5 illustrating a serving mobile network transmitting Stand-alone Dedicated Control Channel (SDCCH) status information to a mobile station in accordance with the teachings of the present invention. A telecommunications node associated with a particular geographic area, such as a cell area, maintains statistical data representing the utilization level of SDCCH logical channels associated with that particular area. Such a telecommunications node may comprise a base station controller (BSC) 60 serving that particular geographic area. It may further comprise a base transceiver station (BTS) serving that particular cell area. Each time a SDCCH channel is requested and allocated by one of the associated base transceiver station (BTS) 30 for a particular mobile station, the

statistical data associated with that particular cell area is updated by an application module 500 associated with the BSC 60. Similarly, each time an allocated SDCCH logical channel is released by a mobile station, the statistical data is updated to reflect the availability of the released channel within the geographic area.

In case the maintained statistical data reflecting the level of SDCCH utilization exceeds a threshold level imposed by the associated mobile telecommunications network, indication is noted that the level of SDCCH utilization within the serving geographic area has reached an undesirable level and there are inadequate amount of available SDCCH logical channels remaining to adequately handle potential high-level mobile service requests from its associated mobile stations. As a result, the application module 500 transmits a broadcast message to all mobile stations located within its coverage area over one of its broadcast channels. In order to reach all mobile stations currently traveling within the effected geographic area, the message may be transmitted over a Broadcast Control Channel (BCCH). The transmitted broadcast message indicates the over-utilization of SDCCH logical channel resources within the current geographic area and instructs the receiving mobile stations to delay requesting low-level mobile service from the serving mobile network.

As another embodiment of the present invention, a plurality of threshold levels may be assigned to the serving mobile telecommunications network. As an illustration, a first threshold level is assigned allowing access to all mobile services except the lowest level mobile services, such as SMS or USSD services. A second threshold level may further be assigned to restrict the next level of mobile service. For example, location updates and IMSI attach may further be restricted from accessing the network. The highest threshold level may then be imposed to allow only access for call connections.

Such a hierarchical structure of mobile services may be identified and determined by the service operator and freely associated with dynamically assignable threshold values. The serving BSC 60 then transmits an appropriate broadcast message informing the associated mobile stations with which level the current utilization level is currently associated.

Upon receiving such SDCCH status information, a mobile station wanting to request mobile service first determines whether the mobile service has higher priority than the received SDCCH status. If the desiring mobile service has lower priority than the current SDCCH resource status, an application module 510 within the mobile station 20 queues the received requests. An indication to the associated mobile subscriber that the requested mobile service is being queued and delayed may further be displayed. Thereafter, the mobile station 20 periodically monitors the BCCH logical channel to determine whether the SDCCH utilization level has been reduced enough to allow access to the mobile network. Such a determination can be made by a number of ways. A different broadcast message may be transmitted by the serving BSC 60 notifying the mobile stations traveling within its effective area that the SDCCH utilization level has decreased. As an alternative, if the monitoring mobile stations no longer receives broadcast messages over the BCCH logical channel, a presumption is then made by the application module 510 that the restriction on the SDCCH channel request is no longer valid. The application module 510 then retrieves the previously queued request and the retrieved mobile service request is then performed in a conventional manner.

Reference is now made to FIG. 6 illustrating the steps performed by a mobile station to comply with the SDCCH resource management procedure in accordance with the teachings of the present invention. A broadcast channel, such as the Broadcast Control Channel (BCCH), is

periodically monitored by the mobile station current located within the particular geographic area at step 500. Thereafter, the mobile station receives an indication to request mobile service towards the serving mobile network. Such an indication may include an associated subscriber entering a directory number, service codes or function keys. Moreover, such an indication may be generated and detected internally within the mobile station. Furthermore, such mobile service may include originating an outgoing call connection, transmitting SMS or USSD messages, or performing location update. In response, an application module associated with the mobile station determines whether a broadcast message associated with the SDCCH resources has been received over the BCCH logical channel. If no such status message has been received, the mobile station takes the "No" decision link 570 and performs the requested mobile service in a conventional manner at step 560. On the other hand, if such status information has been received over the broadcast channel, the mobile station compares the priority of the requested mobile service with the received SDCCH resource status at step 520. If the mobile station is requesting a service higher than the SDCCH resource status currently indicated by the received broadcast message, the mobile station takes the "No" decision link 590 and requests the mobile service in a conventional manner at step 560. If, however, the requesting mobile service has lower priority than the received channel resource status, the application module associated with the mobile station delay transmitting the request by buffering or queuing the received request. Such a request may be queued within a Subscriber Identity Module (SIM) card associated with the mobile station.

Thereafter, the mobile station monitors the broadcast channel to determine whether the channel resource utilization level has decreased enough to enable the mobile station to request the queued mobile service. As

described previously, such a determination can be made either by receiving a different broadcast message indicating a lower utilization level or by not receiving any broadcast message over a predetermined period of time. As an illustration, if the same broadcast message restricting the mobile station from requesting low level mobile service is being transmitted over the BCCH logical channel, the mobile station awaits until the restricting message is no longer received at step 540.

After making a determination that the SDCCH utilization level has decreased, the mobile station then retrieves the previously queued mobile service at step 550. The retrieved mobile service is then requested in a conventional manner at step 560.

Accordingly, by enabling mobile stations to comply with transmitted broadcast messages, the serving mobile telecommunications network is able to better utilize valuable SDCCH logical channel resources for high priority mobile service when the SDCCH utilization has exceed an undesirably high level.

Although a preferred embodiment of the method and apparatus of the present invention has been illustrated in the accompanying Drawings and described in the foregoing Detailed Description, it will be understood that the invention is not limited to the embodiment disclosed, but is capable of numerous rearrangements, modifications and substitutions without departing from the spirit of the invention as set forth and defined by the following claims.

WHAT IS CLAIMED IS:

1. A method for communicating data reflecting the status of channel resources to a plurality of mobile stations within a mobile telecommunications network, said method comprising the steps of:

maintaining a plurality of first channels for facilitating data communication between a telecommunications node serving said plurality of mobile stations and said plurality of mobile stations;

determining that the utilization of said plurality of first channels by some of said plurality of mobile stations has reached a threshold level associated with said telecommunications node; and

transmitting, in response to said determination, first data reflecting the status of said utilization of said plurality of first channels to said plurality of mobile stations over a second channel.

2. The method of claim 1 wherein said first data are transmitted periodically until said utilization of said plurality of first channels has fallen below said threshold level.

3. The method of claim 1 further comprising the steps of:

determining that said utilization of said plurality of first channels has fallen below said threshold level; and

transmitting second data reflecting the availability of said plurality of first channels to said plurality of mobile stations over said second channel.

4. The method of claim 1 wherein said telecommunications node is associated with a plurality of threshold levels and wherein said step of determining further comprises the step of determining with which one of said plurality of threshold levels said utilization of

said plurality of first channel is associated and wherein said transmitted first data further indicates with which one of said plurality of threshold levels said determined status is associated.

5. The method of claim 1 wherein said first data further reflects the scarcity of said plurality of first channels and instructs said plurality of mobile stations to delay requesting low priority mobile services from said telecommunications node.

6. The method of claim 5 wherein said low priority mobile services include a Short Message Service (SMS).

7. The method of claim 5 wherein said low priority mobile services include a Unstructured Supplementary Service Data (USSD) service.

8. The method of claim 1 wherein said plurality of first channels comprise Stand-alone Dedicated Control Channels (SDCCHs).

9. The method of claim 1 wherein said second channel comprises a Broadcast Control Channel.

10. A method for optimizing the utilization of a plurality of first logical channels for a high priority mobile service within a mobile telecommunications network, said method comprising the steps of:

receiving a first indication at a particular mobile station indicating the availability of said plurality of first logical channels within said mobile telecommunications network;

receiving a second indication to request a mobile service towards said telecommunications node at said mobile station;

determining that said requested mobile service has lower priority in comparison to said availability of said plurality of first logical channels; and

delaying said request for said mobile service towards said telecommunications node in response to said determination.

11. The method of claim 10 wherein said first indication comprises data received over a Broadcast Control Channel.

12. The method of claim 10 wherein said first logical channels comprises Stand-alone Dedicated Control Channels (SDCCHs).

13. The method of claim 10 further comprising the steps of:

determining that current availability of said plurality of said first channels within said mobile telecommunications network is now able to handle said requested mobile service; and

requesting said mobile service towards said telecommunications node by requesting one of said plurality of said first logical channels.

14. A system for communicating data indicating the status of channel resources to a plurality of mobile stations within a mobile telecommunications network, said channel resources associated with a plurality of first logical channels for facilitating data communication between said plurality of mobile stations and a telecommunications node serving said plurality of mobile stations, comprising:

an application module for maintaining said plurality of first logical channels;

a processor for determining that the utilization of said plurality of first logical channels by some of said

plurality of mobile stations has exceeded a threshold level associated with said telecommunications node; and

a transmitter for transmitting, in response to said determination, first data informing such determination to said plurality of mobile stations.

15. The system of claim 14 wherein said first data further instructs said plurality of mobile stations to delay requesting low priority mobile services from said telecommunications node.

16. The system of claim 14 wherein said low priority mobile services include a Short Message Service (SMS).

17. The system of claim 14 wherein said low priority mobile services include a Unstructured Supplementary Service Data (USSD).

18. The system of claim 14 wherein said first logical channels comprise a Stand-alone Dedicated Control Channel (SDCCH).

19. The system of claim 14 wherein said first data are transmitted over a Broadcast Control Channel.

20. A mobile station for optimizing the utilization of a plurality of first logical channels for a high priority mobile service within a mobile telecommunications network, said mobile telecommunications network including a telecommunications node providing mobile service to said mobile station, comprising:

means for receiving a first indication indicating the availability of said plurality of first logical channels within said mobile telecommunications network;

means for receiving a second indication to request a mobile service towards said telecommunications node;

means for determining that said requested mobile service has lower priority in comparison to said availability of said plurality of first logical channels; and

means for delaying said request for said mobile service towards said telecommunications node in response to said determination.

21. The mobile station of claim 20 wherein said first indication comprises data received over a Broadcast Control Channel.

22. The mobile station of claim 20 wherein said first logical channels comprises Stand-alone Dedicated Control Channels (SDCCHs).

23. The mobile station of claim 20 further comprising:

means for determining that current availability of said plurality of said first channels within said mobile telecommunications network is now able to handle said requested mobile service; and

means for requesting said mobile service towards said telecommunications node by requesting one of said plurality of said first logical channels.

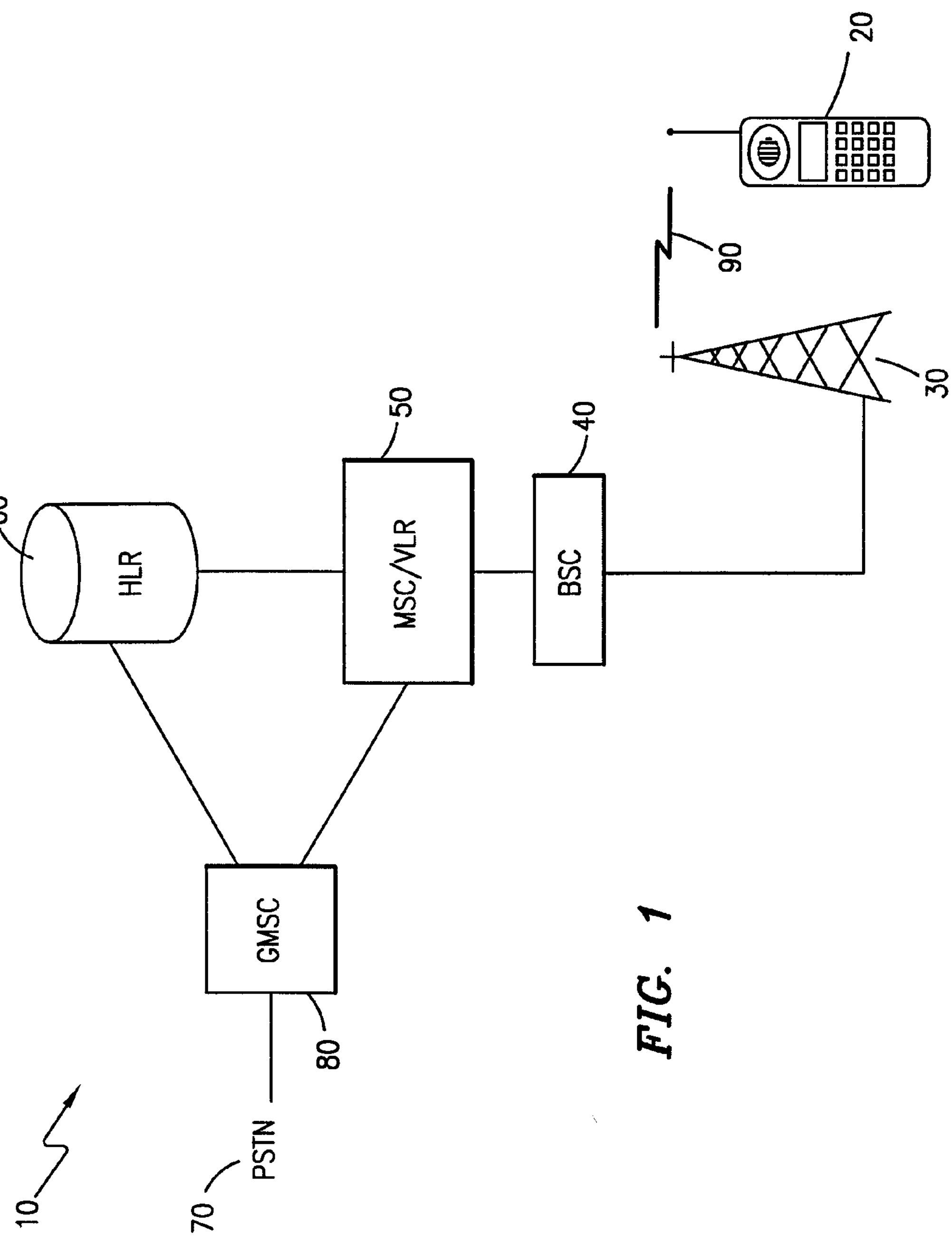

FIG. 1

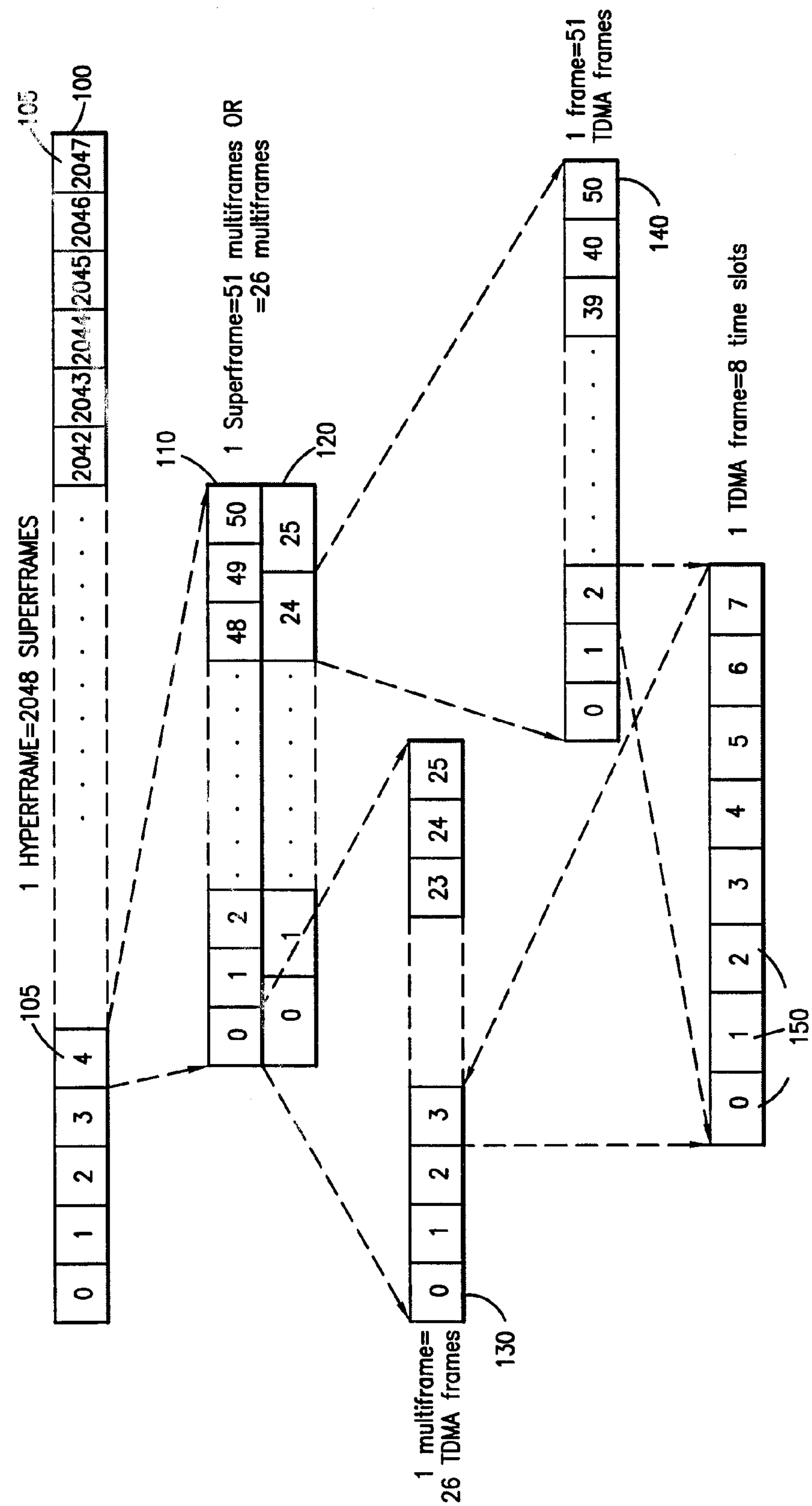
1 HYPERFRAME=2048 SUPERFRAMES
0
1
2
3
4
2042
2043
2044
2045
2046
2047
105
100
1 Superframe=51 multiframes OR
=26 multiframes
48
49
50
24
25
110
120
1 multiframe=
26 TDMA frames
23
130
1 frame=51
TDMA frames
39
40
140
1 TDMA frame=8 time slots
5
6
7
150

FIG. 2

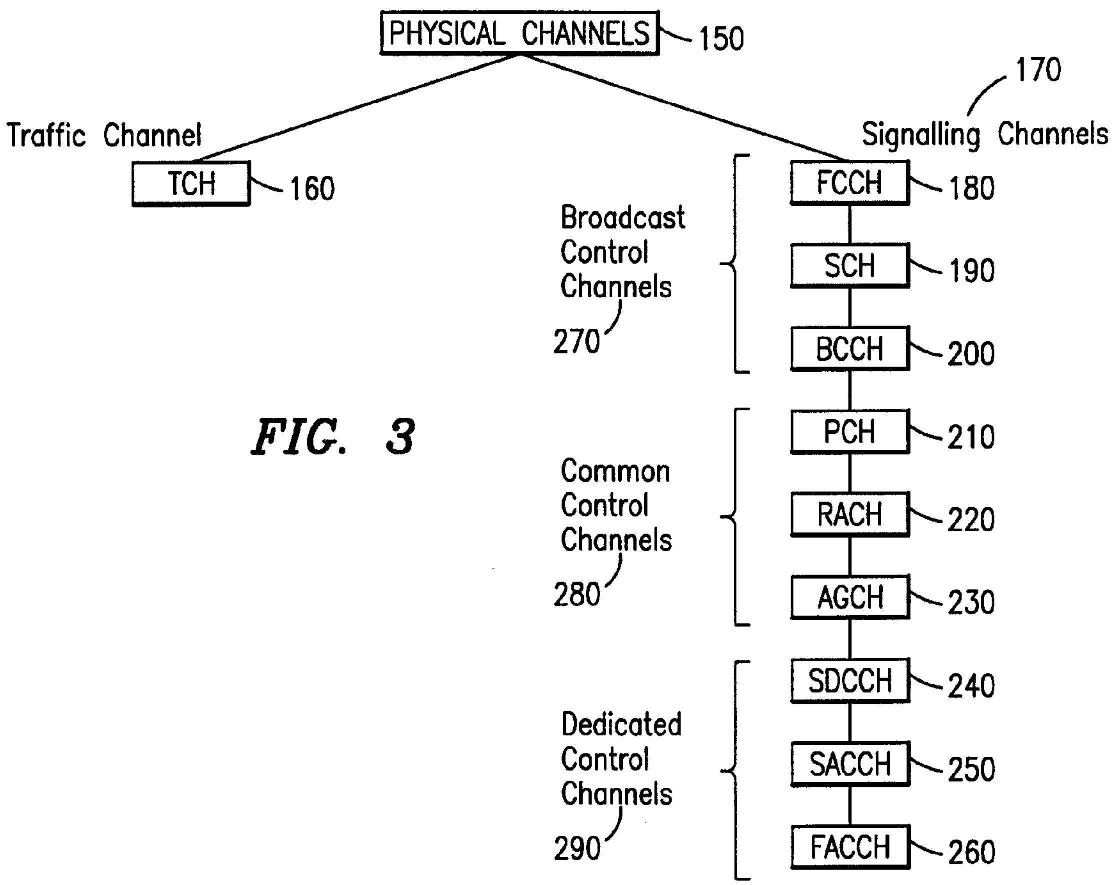
PHYSICAL CHANNELS
150
170
Traffic Channel
Signalling Channels
TCH
160
FCCH
180
Broadcast
Control
Channels
270
SCH
190
BCCH
200
PCH
210
Common
Control
Channels
280
RACH
220
AGCH
230
SDCCH
240
Dedicated
Control
Channels
290
SACCH
250
FACCH
260

FIG. 3

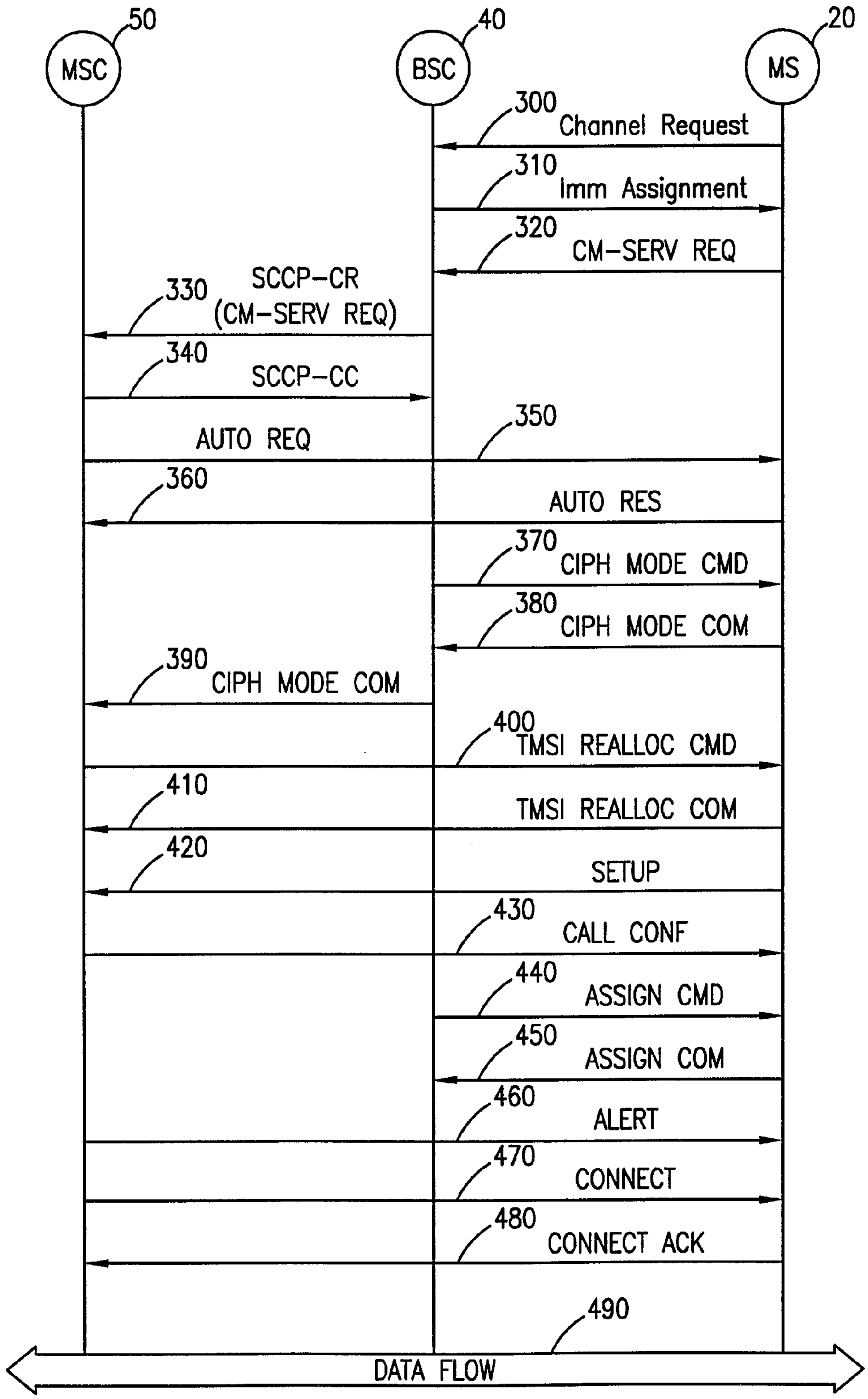
50
MSC
40
BSC
20
MS
300 Channel Request
310 Imm Assignment
320 CM-SERV REQ
330 SCCP-CR
(CM-SERV REQ)
340 SCCP-CC
AUTO REQ
350
360
AUTO RES
370 CIPH MODE CMD
380 CIPH MODE COM
390 CIPH MODE COM
400 TMSI REALLOC CMD
410
TMSI REALLOC COM
420
SETUP
430 CALL CONF
440 ASSIGN CMD
450 ASSIGN COM
460 ALERT
470 CONNECT
480 CONNECT ACK
490
DATA FLOW

FIG. 4

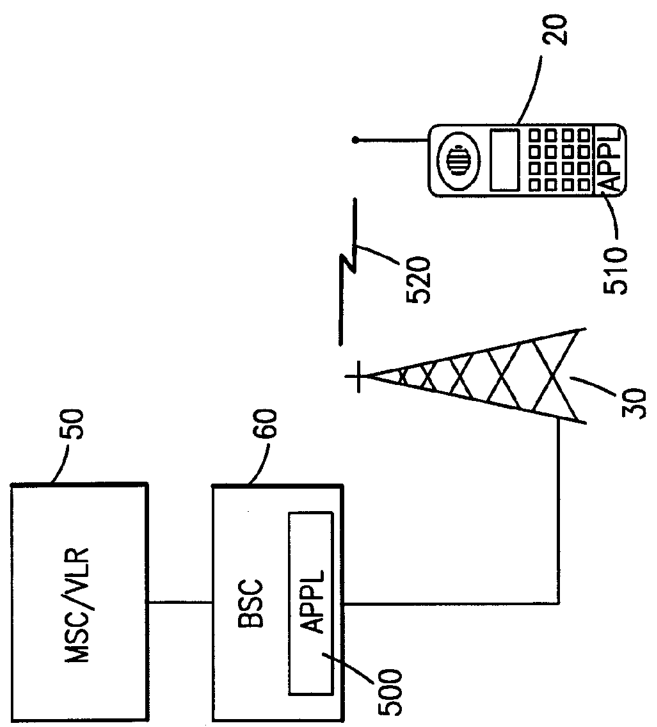

FIG. 5

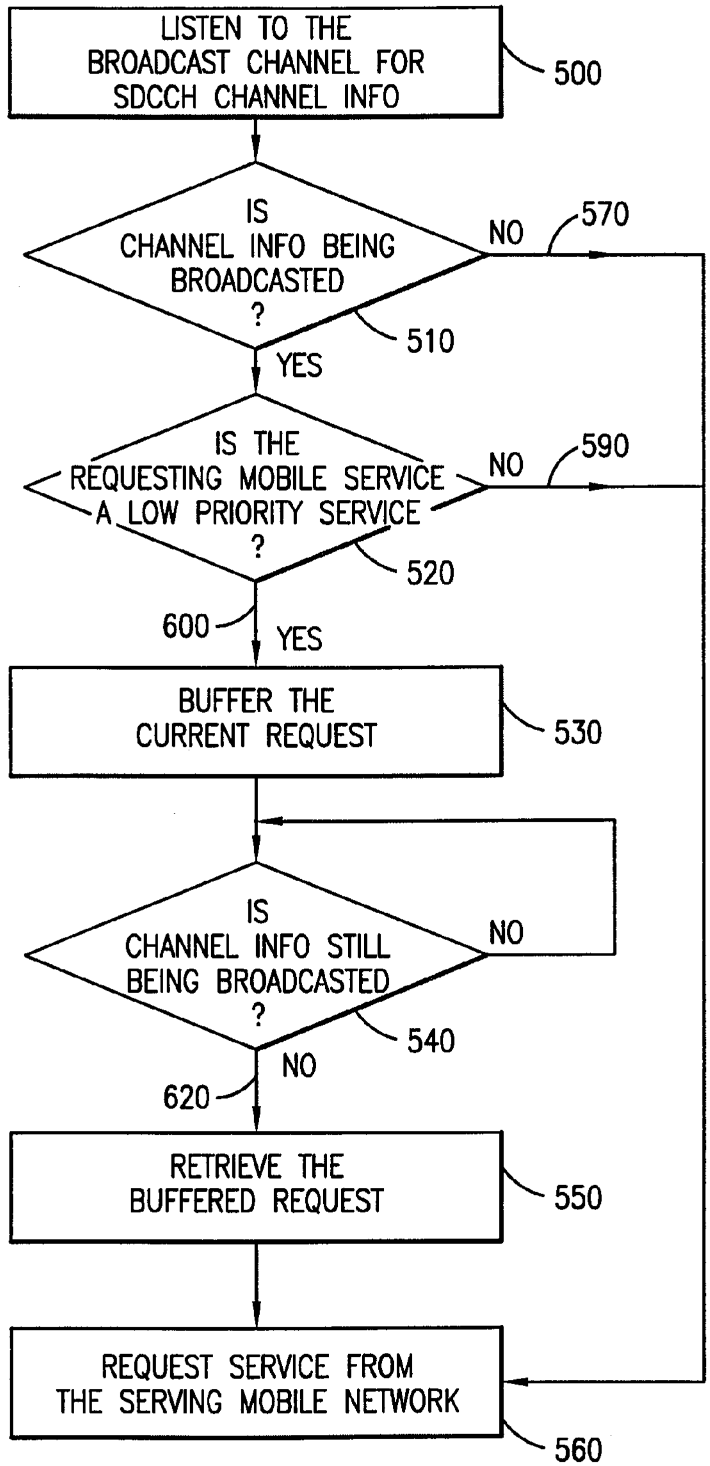
LISTEN TO THE BROADCAST CHANNEL FOR SDCCH CHANNEL INFO
500
IS CHANNEL INFO BEING BROADCASTED ?
NO
570
510
YES
IS THE REQUESTING MOBILE SERVICE A LOW PRIORITY SERVICE ?
NO
590
520
600
YES
BUFFER THE CURRENT REQUEST
530
IS CHANNEL INFO STILL BEING BROADCASTED ?
NO
540
NO
620
RETRIEVE THE BUFFERED REQUEST
550
REQUEST SERVICE FROM THE SERVING MOBILE NETWORK
560

FIG. 6

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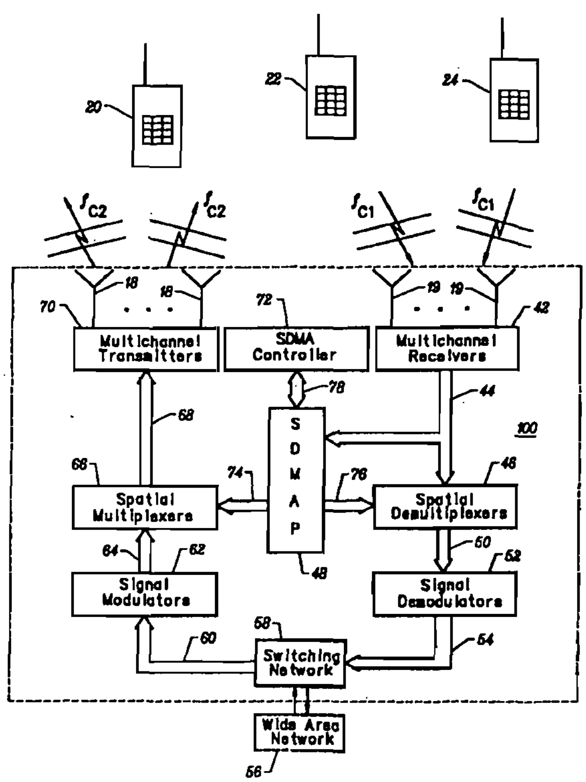
(54) Title: CHANNEL ASSIGNMENT AND CALL ADMISSION CONTROL FOR SPATIAL DIVISION MULTIPLE ACCESS COMMUNICATION SYSTEMS

(57) Abstract

The methods for channel assignment and channel reassignment are suitable for SDMA systems that accommodate the dynamically adaptive spatial channel conditions and allow for more frequent reuse of conventional channels. Three methods for uplink channel assignment are described: a cost function method, a predictive method, and a hierarchical method. The cost function method computes a cost function for each conventional channel based on a weighted correlation matrix for spatial signatures of active subscribers. A spatial channel is created for the selected conventional channel if it is in use. The predictive channel assignment method predicts the uplink received power and interference-plus-noise for each conventional channel. Either the conventional channel with the minimum interference-plus-noise level or the channel with maximal SINR is selected, and a spatial channel is also assigned if the selected channel is in use. The hierarchical method combines the cost function method and the predictive method.



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CHANNEL ASSIGNMENT AND CALL ADMISSION CONTROL FOR SPATIAL DIVISION MULTIPLE ACCESS COMMUNICATION SYSTEMS

CROSS-REFERENCE TO RELATED APPLICATION

This application is a continuation-in-part of pending U.S. patent application entitled "Spectrally Efficient High Capacity Wireless Communication Systems with Spatial-Temporal Processing", Serial No. 08/735,520, filed Oct. 23, 1996.

FIELD OF INVENTION

The present invention relates to wireless communication systems and more specifically to fixed-access or mobile-access wireless networks using spatial division multiple access (SDMA) technology in combination with multiple access systems, such as time domain multiple access (TDMA), frequency division multiple access (FDMA), and/or code division multiple access (CDMA) systems.

BACKGROUND OF THE INVENTION

Wireless communication systems are generally allocated a portion of the radio frequency (RF) spectrum for their operation. The allocated portion of the spectrum is divided into communication channels and channels are distinguished by frequency, time or code assignments, or by some combination of these assignments. Each of these communication channels will be referred to as conventional channels, and a conventional channel will correspond to a full-duplex channel unless otherwise noted. The establishment of a communication link in a communication system depends not only on the availability of a conventional channel but also on the quality of communication that will result from the use of a given available conventional channel.

In wireless communication systems, a conventional channel is used for communication between a base station and a subscriber station. A base station provides coverage to a geographic area known as a cell and may be a point-of presence providing connection between the subscriber station and a wide area network such as a Public

Switched Telephone Network (PSTN). The underlying motivation for the use of cells in wireless systems is the reuse of the RF spectrum in geographically different areas. The reuse of the frequency spectrum can introduce co-channel (intercell) interference between users in different cells that share a common conventional channel. If co-channel interference is not carefully controlled, it can severely degrade the quality of communications. System capacity is in general limited by interference because of the reduction in the number of reusable channels of acceptable quality.

Another source of conventional channel quality degradation is adjacent channel (intracell) interference caused by other conventional channels within a given cell. Ideally, within a given cell each conventional channel should be completely isolated from all of the other conventional channels (orthogonal). In practical systems, full orthogonality between channels can not be ensured because of the complexity and cost such a requirement would place on the system design. For example, adjacent channel interference can result, in FDMA systems, from RF carrier frequency offsets and imperfect filters; in TDMA systems, from timing offset and jitter; and, in CDMA systems, from synchronization inaccuracies or RF multipath propagation.

The more recently introduced SDMA systems (Roy *et al.*, U.S. Pat. No. 5,515,378) allow multiple subscribers within a given cell to simultaneously share the same conventional channel without interfering with one another, and further, allow more frequent reuse of conventional channels within a geographical area covering many cells. SDMA exploits the spatial distribution of subscribers in order to increase usable system capacity. Because subscribers tend to be distributed over a cell area, each subscriber-base station pair will tend to have a unique spatial signature characterizing how the base station antenna array receives signals from the subscriber station, and a second spatial signature characterizing how the base station antenna array transmits signals to the subscriber station. Subscribers sharing the same conventional channel are said to be using different spatial channels. As in the case of FDMA, TDMA, and CDMA systems, spatial channels in a SDMA system may not be perfectly orthogonal because of hardware limitations and multipath propagation. It should be noted that non-spatial multiplexing (*e.g.*, FDMA, TDMA, and CDMA), when used in combination with antenna array patterns that are controlled by using spatial processing, is referred to as

SDMA in the context of this invention. In practice, spatial signatures and antenna arrays can be used in a non-spatial-division-multiple-access system configuration for enhancing communications between base stations and subscribers by use of spatial signal processing techniques. In these cases, the label SDMA will still be used in the context of the description that follows.

Fig. 1 shows an example of a wireless SDMA TD/FD/CDMA system (Barratt *et al.*, U.S. Patent Application No. 08/375,848) in which a number of subscriber stations (symbolically shown as handsets) 20, 22, 24 are being served by base station 100 that may be connected to a wide area network (WAN) 56 for providing any required data services and connections external to the immediate wireless system. Switching network 58 interfaces with WAN 56 for providing multichannel duplex operation with the WAN by switching incoming WAN data to lines 60 of base station 100 and switching outgoing signals from base station 100, on line 54 to the WAN. Incoming lines 60 are applied to signal modulators 62 that produce modulated signals 64 for each subscriber station the base station is transmitting to. A set of spatial multiplexing weights 74 for each subscriber station are applied to the respective modulated signals in spatial multiplexers 66 to produce spatially multiplexed signals 68 to be transmitted by a bank of multichannel transmitters 70 using transmit antenna array 18. The SDMA processor (SDMAP) 48 produces and maintains spatial signatures for each subscriber station for each conventional channel, calculates spatial multiplexing and demultiplexing weights for use by spatial multiplexers 66 and spatial demultiplexers 46, and uses the received signal measurements 44 to select a channel for a new connection. In this manner the signals from the current active subscriber stations, some of which may be active on the same conventional channel, are separated and interference and noise suppressed. When communicating from the base station to the subscriber stations, an optimized multilobe antenna radiation pattern tailored to the current active subscriber station connections and interference situation is created. An example of a transmit antenna pattern that may be created is shown in Fig. 2.

Returning to Fig. 1, spatial demultiplexers 46 combine received signal measurements 44 from the multichannel receivers 42 and associated antenna array 19 according to spatial demultiplexing weights 76, a separate set of demultiplexing weights

being applied for each subscriber station communicating with the base station. The outputs of spatial demultiplexers 46 are spatially separated signals 50 for each subscriber station communicating with the base station, which are applied to signal demodulators 52 to produced demodulated received signals 54 for each subscriber station communicating with the base station. In an alternate embodiment, the demultiplexing and demodulation processing are performed together in a nonlinear multidimensional signal processing unit. The demodulated received signals 54 are then available to switching network 58 and WAN 56.

In an FDMA system implementation, each multichannel receiver and each multichannel transmitter is capable of handling multiple frequency channels. In other embodiments, multichannel receivers 42 and multichannel transmitters 70 may instead handle multiple time slots, as in a TDMA system; multiple codes, as in a CDMA system, or some combination of these well known multiple access techniques (Barratt *et al.*, U.S. Patent Application No. 08/375,848).

In practical systems that may involve hundreds or thousands of subscriber stations, perfect separation or orthogonality between every subscriber station, following the application of SDMA processing, cannot be insured because of the complexity and cost that such a requirement would place on the system design. If the separation of subscriber station connections post-SDMA processing cannot be ensured, the extended capacity of the SDMA will be limited and interference between subscribers will occur from the use of SDMA techniques. The consequence of this practical limitation is that a method for minimizing the interference and thereby maximizing the effective channel capacity of the SDMA system is required.

Even if two or more subscriber stations are not perfectly separated or orthogonal after SDMA processing, it still may be possible to share a common conventional channel in a TDMA, FDMA or CDMA system using SDMA technology. From a practical point of view, it is not required that the subscriber stations be perfectly separated after SDMA processing to share a common conventional channel. It is only required that the interference between subscribers sharing a common conventional

channel post-SDMA processing be low enough so as not to reduce the quality of communications below a prescribed level.

Because of the interference introduced by frequency reuse and the fragile nature of orthogonality for conventional and spatial channels, all wireless multiple access communications systems need a method for base station and channel assignment that minimizes these adverse effects when a new call or connection between a base station and a subscriber is made. The labels new subscriber and new connection will be used interchangeably to denote a new call or connection between a base station and a subscriber station, and the labels active subscriber, existing connection and existing subscriber will be used interchangeably to denote a call or connection in-progress between a base station and a subscriber station. If not careful, the new subscriber may be assigned to a base station and a channel on which poor quality is experienced due to excessive interference. Moreover, the addition of a new subscriber has the potential consequence of adversely affecting the quality of communications on existing connections. Also, existing subscribers can suffer from increased channel interference from the addition of a new subscriber, or other unrelated causes, that can require moving subscribers from currently assigned channels to new channels in order to restore acceptable quality communications. Channel re-assignment methods, using decision processes similar to those used for initial base station and channel assignment, are also required.

Prior art channel assignment and reassignment methods are based on measurements of physical phenomena such as the received signal strength indication (RSSI) or the co-channel interference on different conventional channels. Barnett, in U.S. Patent No. 5,557,657, describes a method for handover between an overlay cell and an underlay cell depending on the RSSI. Booth (U.S. Patent No. 5,555,445) describes a method for intercell handoff in which an intracell handoff from one conventional channel to another is first attempted, and the success or failure of the handoff is indicated by the RSSI. Knudsen (U.S. Patent No. 5,448,621) describes a method for reallocating conventional channels between cells that depends on the number of unused conventional channels in each cell (*i.e.*, the cell load). Grube *et al.* (U.S. Patent No. 5,319,796) outlines a method for measuring co-channel interference on a conventional

channel by placing additional receivers in the coverage area of the co-channel user and then transmitting feedback information on measured co-channel interference to a channel assignment controller. In all of these methods, the processes of channel assignment and reassignment do not take the spatial distribution of the subscribers into account, nor do they consider how the RSSI and co-channel interference jointly affect the signal quality of the new connection.

Hanabe (U.S. Patent No. 5,475,864) describes a channel assignment method for sectorized cells which have static antenna beam patterns. Hanabe does not consider what happens with fully adaptive SDMA systems in which beam patterns dynamically change depending on which subscribers are active at any given time. Furthermore, the channel assignment of spatial channels made possible by SDMA is never addressed.

If two subscribers with similar spatial signatures were to be assigned to the same conventional channel, either at the same base station or at two different base stations, serious interference would render the channel unusable to both subscribers. Hence, there is a need for a new method of channel assignment for advanced, fully adaptive SDMA systems that can predict, a priori, the quality of a spatial or conventional channel; *i.e.*, before the new connection is assigned to a given base station and channel. Also, there is a need for a SDMA channel assignment method that can predict the impact of a new connection on existing connections and can perform call admission control as necessary. The availability of such base station and channel assignment, reassignment, and admission control methods would allow SDMA methods to increase system capacity by better isolating subscribers while maintaining acceptable communications quality.

SUMMARY OF THE INVENTION

The present invention includes methods for channel assignment, channel reassignment, and call admission control suitable for SDMA systems that accommodate the dynamically adaptive spatial channel conditions and allow for more frequent reuse of conventional channels.

Three methods for uplink channel assignment are described: a Weighted Correlation method, a Predicted Quality method, and a Hierarchical method.

The Weighted Correlation method computes a cost function for each conventional channel based on a weighted correlation of the new subscriber's spatial signature with the spatial signatures of the active subscribers. The spatial signatures of the active subscribers need not be explicitly known to compute the cost function. In an embodiment in which no knowledge about the active subscriber signatures is assumed, the cost function is formed using an unstructured estimate of the sample covariance matrix, which is computed from measurements of the antenna array response for a prescribed number of time samples. In an alternate embodiment, a structured estimate of the sample covariance matrix, based on the spatial signatures of active subscribers and a noise-plus-interference covariance matrix, is used to compute the cost function. In one embodiment, a conventional channel with acceptably low cost is assigned to the new connection.. In another embodiment, the conventional channel with the minimum cost is assigned to the new connection. The set of candidate conventional channels from which a channel for assignment is to be selected may be constrained to the subset of channels for which the cost of the new subscriber is less than a prescribed threshold. The assignment of more than one subscriber to the same conventional channel is permitted if there are sufficient hardware resources at the chosen base station for the selected channel to accommodate the new connection. If no candidate channels are found, the new subscriber is not assigned to the chosen base station.

The Predicted Quality channel assignment method predicts the uplink received signal power and interference-plus-noise for each conventional channel, based on an estimate (predicted value) of the sample covariance matrix of received signals at the base station antenna array that may result should the new subscriber be assigned to, and become active on, a given channel. The method may use an unstructured estimate of the sample covariance matrix of received signals from subscribers already active, by measuring the base station array response for a prescribed number of samples, or else the method may use a structured estimate of the sample covariance matrix, based on the spatial signatures of already active subscribers and a noise-plus-interference covariance matrix. In one embodiment, the cost function for a conventional channel is computed as

the predicted interference on that channel. In an alternative embodiment, the cost function for a conventional channel is computed as the negative of the predicted signal-to-interference-plus-noise ratio (SINR) of the new connection on that channel. In either case, the new subscriber may be assigned to a conventional channel with an acceptably low cost, or else to the conventional channel with the minimum cost. The set of candidate conventional channels from which a channel for assignment is to be selected may be constrained to the subset of channels for which the predicted SINR of the new subscriber, and, optionally, the predicted SINRs of the active co-channel subscribers, is/are greater than a prescribed threshold. In all these cases, the assignment of more than one subscriber to the same conventional channel is permitted if there are sufficient hardware resources at the selected base station for the chosen channel to accommodate the new connection. If no candidate channels are found, the new subscriber is not assigned to the selected base station.

The Hierarchical method uses the Weighted Correlation method to select a subset of candidate channels that represent the channels with the lowest cost as determined by the Weighted Correlation method. The channel assignment is then made by applying the Predictive Quality method to the subset of candidate channels.

The Downlink Predictive channel assignment method, when not constrained by the uplink channel assignment, assigns a conventional channel to a new connection by having the new subscriber report the downlink received signal level for each conventional channel and estimating the downlink interference-plus-noise level from the subscriber report. In one embodiment, the cost function for a conventional channel is computed as the downlink interference-plus-noise level on that channel. Alternatively, the downlink spatial signature and associated multiplexing weights of the new connection on each conventional channel are further used to compute a predicted downlink received signal level. The cost function for a conventional channel is then computed as the negative of the predicted downlink SINR for the channel. In either case, the new subscriber may be assigned to a conventional channel with an acceptably low cost, or else to the conventional channel with the minimum cost. The set of candidate conventional channels to select from may be constrained to the subset of channels for which the predicted SINR of the new subscriber, and optionally, the predicted SINRs of

the active co-channel subscribers, is/are greater than a prescribed threshold. In all these cases, the assignment of more than one subscriber to the same conventional channel is permitted if there are sufficient hardware resources at the selected base station for the chosen channel to accommodate the new connection. If no candidate channels are found, the new subscriber is not assigned to the selected base station.

In addition, for all uplink and downlink channel assignment methods an optional distortion criterion may be added which estimates the transmitter and/or receiver distortion effects produced by any particular conventional channel assignment by computing a crest factor. In one embodiment, the effect is expressed by augmenting the cost function for any particular channel assignment method with the crest factor cost. In the preferred embodiment, the effect is expressed by constraining the selection of a conventional channel to among those channels with an acceptably low crest factor.

The Joint Uplink-Downlink channel assignment method combines the cost function of an uplink method with that of a downlink method to form a joint cost function. The new subscriber may be assigned to a conventional channel with an acceptably low joint cost, or else to the conventional channel with the minimum joint cost. The set of candidate conventional channels to select from may be constrained to the subset of channels satisfying the constraints of the uplink method and the downlink method. The assignment of more than one subscriber to the same conventional channel is permitted if there are sufficient hardware resources at the selected base station for the chosen channel to accommodate the new connection. If no candidate channels are found, the new subscriber is not assigned to the selected base station.

All of the channel assignment methods may be applied to either a set of candidate conventional channels associated with any particular base station or else a set of candidate conventional channels associated with a multiplicity of base stations. In the latter embodiment, the channel assignment method performs the selection of a base station in addition to the selection of a conventional channel for the new connection.

Channel re-assignment may be accomplished by any of the above methods for channel assignment, with the modification that the conventional channel the

subscriber wishes to be reassigned from is omitted from the set of candidate conventional channels.

Three methods for call admission control include: a Weighted Correlation method that includes comparing the cost of assigning a selected conventional channel to a new connection against a prescribed threshold, assigning the selected channel if the threshold is exceeded, otherwise rejecting the selected channel; a Predictive method that includes comparing the predicted uplink and/or downlink SINR of a selected conventional channel against the corresponding prescribed uplink and/or SINR threshold(s), assigning the selected channel if the candidate SINR(s) exceed(s) the threshold(s), and blocking the assignment if otherwise; and a general Load Estimation method that is applicable to SDMA and non-SDMA systems that includes estimating the system call load, prescribing a call load threshold indicative of the number of calls in progress, comparing the estimated system call load with the call load threshold, assigning the selected channel if the estimated load is less than threshold, and blocking assignment if otherwise.

BRIEF DESCRIPTION OF THE DRAWINGS

Fig. 1 is a graphical representation of a SDMA system;

Fig. 2 is a graphical representation of the SDMA multichannel transmitters' antenna patterns generated from three multiplexing weight vectors;

Fig. 3(a) is a flow diagram of the weighted correlation method for channel assignment;

Fig. 3(b) is a flow diagram for computing the unstructured form of the covariance matrix;

Fig. 3(c) is a flow diagram for computing the structured form of the covariance matrix;

Fig. 4 is a flow diagram of the predicted quality method for channel assignment;

Fig. 5 is a flow diagram for the hierarchical method for channel assignment;

Fig. 6 is a flow diagram for the downlink predictive channel assignment method; and

Fig. 7 is a flow diagram for three methods of call admission control.

DETAILED DESCRIPTION OF THE INVENTION

Channel assignment in a full-duplex communication channel includes the selection of both an uplink channel (from subscriber to cell base station) and a downlink channel (from cell base station to subscriber). The case of half-duplex channel assignment may be considered as a special case of the full-duplex problem. Interference on the uplink channel comes primarily from other subscriber stations while interference on the downlink channel is caused primarily by base stations of other cells. Consequently, the quality of communications on the uplink and downlink channels will generally differ. In one embodiment of the invention, uplink and downlink channel assignments are performed independently and separately and, because of this lack of constraints in selecting the uplink and downlink channels, offers the potential for achieving the highest system capacity. However, many practical systems impose a fixed relationship between the uplink and downlink channels so that independent selection is not possible.

For example, in the Personal Handyphone System (PHS) standard (Association of Radio Industries and Businesses (ARIB) Preliminary Standard, Version 2, RCR STD 28, approved by the Standard Assembly Meeting of December, 1995), the uplink and downlink channels form a full-duplex channel and must be on the same RF carrier, so that the carrier frequency of uplink and downlink channel can not be independently specified. Also, the downlink time division multiplexed time-slot is specified as preceding the uplink time-slot by exactly four time-slots. For such systems, the selection of either uplink or downlink channel automatically determines the selection of the other. In one embodiment, the selection of a full-duplex channel is achieved by performing an uplink channel assignment and specifying the downlink assignment in accordance with the existing rules of the system. This method is advantageous when the capacity of the full-duplex channel system is primarily uplink channel capacity limited. In another embodiment, the full-duplex channel is chosen by performing the downlink channel assignment and allowing the choice of the uplink assignment to be fixed by the rules of the system. This method is preferred when the system is primarily downlink

channel capacity limited. In another embodiment, the assignment of the uplink and downlink channel is considered jointly by evaluating each uplink-downlink channel pair as a unit and assigning a new subscriber to the best uplink-downlink pair. This method is preferred for systems in which the channel capacity is dominated by neither uplink nor downlink channel capacity.

Other practical considerations for channel assignment are the dynamic range of the RF power amplifiers (PAs) and whether the base station transmitter architecture is wideband or narrowband. The dynamic range of a PA is limited on the low power end by the noise floor and on the high power end by the maximum PA output for which the distortion of the amplified signal remains acceptably low. A channel assignment method must be aware of the PA dynamic range characteristics when selecting a channel because the required transmit power may differ from channel to channel.

A base station transmitter PA also generates intermodulation distortion from the mixing of RF subcarriers of differing frequencies. In a narrowband RF transmitter architecture, the power delivered for each subcarrier is provided by a separate PA so that the mixing of different subcarrier bands does not occur, greatly reducing the intermodulation distortion. Further, any distortion products generated from in-band mixing by each PA which fall outside of the subcarrier band can be filtered so as to minimize the distortion products that can cause interference with other subcarriers. By contrast, a transmitter with a wideband PA architecture uses a multi-carrier power amplifier (MCPA) that amplifies a group of subcarriers simultaneously, producing intermodulation distortion from the mixing of different subcarriers. The intermodulation distortion so generated overlaps with the group of subcarrier bands carrying the subscriber signals and cannot be separated and filtered. The MCPA produces intermodulation products due to the presence of a multicarrier RF signal. While it is possible to produce an MCPA with very low intermodulation distortion, the cost of doing so is very high. Thus, there is a need for a channel assignment method that helps mitigate the effects of intermodulation distortion by taking the PA architecture into account and permitting a lower cost solution to be used.

Analogous problems with distortion also exist in a wideband receiver architecture. For example, sources of non-linearities in the receiver include RF mixers, low noise amplifiers and analog-to-digital converters. A channel assignment algorithm must take this information into account or else the capacity of the network may be adversely affected.

As previously explained, in SDMA there are two spatial signatures associated with each subscriber-base station pair on a particular conventional channel (Barratt *et al.*, U.S. Patent Application No. 08/375,848). A base station associates with each subscriber station a receive, or uplink, spatial signature related to how that subscriber station receives signals transmitted to it by the base station's antenna array and a transmit, or downlink, spatial signature related to how the base station's receive antenna array receives signals transmitted by the subscriber station. The transmit and receive spatial signatures contain information about the amplitude attenuation and relative phase of the RF signal at each antenna element transmitter and receiver, respectively, of the base station. This amplitude and phase information at each receiver or transmitter can be treated as vector elements, $\{a_i\}$, of a complex column vector, a. The spatial signatures can be stored in a database and updated at prescribed intervals, or they may be estimated during the initial phase of a call setup when a new connection from a subscriber is initiated, or they may be analytically determined (Roy *et al.*, U.S. patent 5,515,378). For example, in the case of PHS, a link channel establishment phase takes place on the signaling control channel (SCCH) before communicating on an assigned link (traffic) channel (LCH). During this link channel establishment phase, the spatial signatures of the new subscriber can be measured.

The spatial signatures contain information about the ability to communicate with a subscriber. If a_k^i and a_k^j are the receive spatial signatures for subscribers i and j, respectively, on conventional channel k, then their normalized absolute inner product is defined as $\frac{\left|a_k^{i*} a_k^j\right|}{\left\|a_k^i\right\| \cdot \left\|a_k^j\right\|}$, where $|\ |$ denotes the complex modulus, $(\ )^*$ denotes the complex conjugate transpose and $\|\ \|$ denotes the Euclidean norm of a complex vector. The normalized absolute inner product of a_k^i and a_k^j is indicative of

the ability to simultaneously communicate to subscribers i and j on the same uplink conventional channel. Orthogonal signatures would have a normalized absolute inner product of zero, indicating that interference between subscribers is unlikely even if both share a common conventional channel. A significant normalized absolute inner product value would indicate a potential interference problem if both subscribers were to share a common conventional channel. However, there are two problems with the use of the normalized absolute inner product as a basis of channel assignment: it may be too difficult (*i.e.*, complex and/or expensive) to keep track of the spatial signatures for all channels of all subscribers in adjacent cells; and a significant normalized absolute inner product does not necessarily indicate an interference problem because, for example, subscribers within different cells may have a large normalized absolute inner product between spatial signatures and not interfere if they are isolated by large distances or by high loss RF propagation paths. Therefore, the received signal levels of the subscribers on a given conventional channel from surrounding subscribers will also determine whether there will be an unacceptable level of interference on that channel.

Several optional approaches to uplink channel assignment are available, each varying in relative complexity and performance characteristics: a Weighted Correlation method, a Predicted Quality method, and a Hierarchical method combining both the Weighted Correlation method and the Predicted Quality method.

The Weighted Correlation method defines a quadratic cost function for the k^{th} conventional channel as

$$c_k = a_k^* \hat{R}_{zz}^{(k)} a_k \qquad \textbf{Eq. 1}$$

where a_k is the uplink spatial signature of the new subscriber on conventional channel k and $\hat{R}_{zz}^{(k)}$ is the sample covariance matrix of the antenna array response of conventional channel k. The spatial signature a_k is typically estimated during call setup, or it may be stored in a database and updated at prescribed intervals. An unstructured estimate of the sample covariance matrix which does not require a priori knowledge of the spatial signatures of the active subscribers may be computed, typically by measuring the

received signal at each antenna element receiver of Fig. 1 over several time samples and averaging; *i.e.*,

$$\hat{R}_{zz}^{(k)} = \frac{1}{n}\sum_{i=1}^{n} z^{(k)}(i) z^{(k)*}(i) \qquad \textbf{Eq. 2}$$

where $z^{(k)}(i)$ is the vector received signal of the antenna array on conventional channel k at time i, and n is the number of time samples. For example, in the PHS system, n may be chosen to be the number of data symbols in a PHS burst. In one embodiment, the new subscriber's call is assigned to a conventional channel with an acceptable cost, c_k. The acceptable cost level can be that which corresponds to an acceptable bit error rate for communication between the base station and the subscriber station. In the preferred embodiment, the new subscriber's call is assigned to the conventional channel for which the cost function, c_k, of Eq. (1) is minimal. The assignment of more than one subscriber to the same conventional channel is permitted if there are sufficient hardware resources at the selected base station for the chosen channel to accommodate the new connection.

If an unstructured estimate of $\widehat{R}_{zz}^{(k)}$ is used in Eq. (1), then the channel assignment method described above can operate in the absence of any form of information exchange or communication between different base stations. By performing an unstructured estimate, the spatial signatures of all active subscribers, both within the same cell as well as in neighboring cells, have been accounted for without having to explicitly measure them one-by-one.

Alternatively, a structured estimate of the sample covariance matrix which takes advantage of any knowledge about the spatial signatures of active subscribers may be computed. In a wireless system employing SDMA, a base station may know the spatial signatures and transmit signal powers of the active subscribers with which it is communicating. Hence, in an alternative embodiment, the cost function of Eq. (1) can be computed by performing a structured estimate of the sample covariance matrix $\widehat{R}_{zz}^{(k)}$:

$$\widehat{R}_{zz}^{(k)} = A_k R_{ss}^{(k)} A_k^{*} + R_{nn}^{(k)} \qquad \textbf{Eq. 3}$$

where A_k is the collection of spatial signatures formed by column-wise concatenation of the spatial signatures on conventional channel k of active subscribers in the same cell as the new connection, $R_{ss}^{(k)}$ is a an expected cross-correlation matrix whose diagonal elements are the average transmit powers of the active subscribers on conventional channel k in the same cell as the new connection, and $R_{nn}^{(k)}$ is a noise-plus-interference covariance matrix containing the contributions of noise and intercell interference to the received signals at the base station antenna array. $R_{nn}^{(k)}$ may be estimated by measuring the received signal at each antenna element receiver of Fig. 1 during time intervals when the active subscribers on channel k and in the same cell as the new connection are not transmitting and then time averaging, as in Eq. (2). Alternatively, if the spatial signatures and transmit powers of all active subscribers are available at each base station, $R_{nn}^{(k)}$ may be estimated as

$$R_{nn}^{(k)} = A_k^0 R_{s_0 s_0}^{(k)} A_k^{0*} + Q \qquad \textbf{Eq. 4}$$

where $R_{s_0 s_0}^{(k)}$ is an expected cross-correlation matrix whose diagonal elements are the average transmit powers of the active subscribers on conventional channel k in cells different from that of the new connection. A_k^0 is the collection of spatial signatures formed by column-wise concatenation of the known spatial signatures on conventional channel k of active subscribers in cells different from that of the new connection received at the base station of the new connection. Q is the estimated receiver noise covariance matrix. The contents of Q can also be regarded as regularization parameters chosen by the user. In many common cases, $Q = \sigma^2 I$, where I is the identity matrix and σ^2 is the estimated receiver noise.

In one embodiment, the new subscriber's call is assigned to a conventional channel with an acceptable cost, c_k. In the preferred embodiment, the new subscriber's call is assigned to the conventional channel for which the cost function, c_k, of Eq. (1) is minimal. The assignment of more than one subscriber to the same conventional channel is permitted if there are sufficient hardware resources at the selected base station for the chosen channel to accommodate the new connection in additional to any existing connections.

In a typical SDMA system, the sample covariance matrix of Eq. (2) or Eq. (3) is computed and continually updated as part of the spatial processing for each conventional channel, thus obviating the need to recompute it for channel assignment. The computations required by Eq. (1) are then minimal.

A further improvement in channel assignment is obtained by use of the Predicted Quality Channel Assignment Method by predicting the quality of communication that will result from assigning a new connection to a particular conventional channel. This is accomplished by estimating the signal power and the interference-plus-noise power that a new subscriber will experience on each conventional channel if assigned to that channel by using a model of the RF environment and the SDMA processing, without actually assigning the call to any conventional channel.

Let $\hat{R}_{zz}^{(k)}$ represent the sample covariance matrix before the new subscriber is assigned to conventional channel k, and $\tilde{R}_{zz}^{(k)}$ represent the predicted covariance matrix if the new subscriber were to be assigned to and become active on conventional channel k. As described previously, $\hat{R}_{zz}^{(k)}$ may be computed by either an unstructured estimate (Eq. (2)) or a structured estimate (Eq. (3)). In the preferred embodiment, the relationship between $\hat{R}_{zz}^{(k)}$ and $\tilde{R}_{zz}^{(k)}$ is modeled by

$$\tilde{R}_{zz}^{(k)} = \hat{R}_{zz}^{(k)} + a_k r_{ss}^{(k)} a_k^* \qquad \textbf{Eq. 5}$$

where a_k is the uplink spatial signature of the new subscriber on conventional channel k and $r_{ss}^{(k)}$ is a scalar quantity representing the transmit power of the new subscriber on conventional channel k.

In the preferred embodiment, the uplink spatial demultiplexing weights, $\tilde{w}_k^U$, for the new subscriber on conventional channel k are expressed as the column vector

$$\tilde{w}_k^U = (\tilde{R}_{zz}^{(k)})^{-1} r_{ss} a_k \qquad \textbf{Eq. 6}$$

It should be noted that the above expression requires the inverse of $\tilde{R}_{zz}^{(k)}$ which is usually undesirable because computing the inverse is computationally expensive. However, by taking advantage of the model for $\tilde{R}_{zz}^{(k)}$ in Eq. (5), and by invoking the Sherman-Morrison inversion formula ("Matrix Computations", Golub *et al.*, The Johns Hopkins University Press, Baltimore, MD, 1983, p. 3), the inverse of $\tilde{R}_{zz}^{(k)}$ may be expressed as

$$(\tilde{R}_{zz}^{(k)})^{-1} = (\hat{R}_{zz}^{(k)})^{-1} - (\hat{R}_{zz}^{(k)})^{-1} a_k a_k^* (\hat{R}_{zz}^{(k)})^{-1} / ((1/r_{ss}^{(k)}) + a_k^* (\hat{R}_{zz}^{(k)})^{-1} a_k) \qquad \textbf{Eq. 7}$$

Thus, by using this expression the process of matrix inversion has been reduced to a series of simpler matrix multiplication computations. Further, it should be noted that in a typical SDMA system, the inverse of the sample covariance matrix before the new subscriber is assigned to conventional channel k, $(\hat{R}_{zz}^{(k)})^{-1}$, has been computed and is continually updated as part of the spatial processing for the already active subscribers on each conventional channel k, making it unnecessary to compute it when using Eq. (7).

The predicted uplink signal power that would result from the assignment of the new subscriber to channel k, S_k^U, is estimated as

$$S_k^U = \left|\tilde{w}_k^{U*} a_k\right|^2 r_{ss} \qquad \textbf{Eq. 8}$$

and I_k^U, the uplink interference-plus-noise power for the new subscriber, is estimated as

$$I_k^U = \tilde{w}_k^{U*} \hat{R}_{zz}^{(k)} \tilde{w}_k^U \qquad \textbf{Eq. 9}$$

Having computed S_k^U and I_k^U for each conventional channel k, the uplink signal-to-interference-plus-noise ratio (SINR) of the new connection on each channel k is estimated by

$$SINR_k = S_k^U / I_k^U \qquad \textbf{Eq. 10}$$

In one embodiment, the cost function for conventional channel k is computed as I_k^U. The new subscriber is assigned either to the first conventional channel for which the computed cost is acceptably low, or else to the conventional channel with the minimal

cost. (One definition for an acceptably low computed cost is a cost that is equal to, or less than, the cost that corresponds to a maximal acceptable bit error rate between a subscriber and the base station.) The rationale is that a conventional channel assigned to the new subscriber on which interference from existing subscribers is acceptably low will, by reciprocity, tend to be a channel on which the new subscriber will produce acceptably low interference with the existing subscribers.

In an alternative embodiment, the cost function for any conventional channel k is computed as $-SINR_k$, the negative of the predicted SINR on that channel. The new subscriber is assigned either to the first conventional channel for which the computed cost function is acceptably low, or else to the conventional channel with the minimal cost. This method is useful for, but not limited to, wireless systems which employ means for controlling transmit power levels as are well-known in the art. In this way, the new subscriber can use the lowest transmit power and thereby maximally reduce interference in the system. However, for various reasons (*e.g.,* if the power control range is very limited), conventional channel assignment can always be made under the rule described in the previous paragraph for Predicted Quality channel assignment.

For either of the embodiments described in the preceding two paragraphs, the set of candidate channels that the channel assignment method chooses from may be constrained to the subset of channels for which the predicted signal-to-interference-plus-noise-ratios of the new connection exceed a prescribed threshold level. Typically the threshold is set at or near the SINR required to maintain an acceptable bit-error-rate for connection between the subscriber station and the base-station.

In a wireless system employing SDMA, a base station typically knows the spatial signatures and transmit powers of the active subscribers with which it is communicating. This knowledge about the co-channel active subscribers may optionally be exploited in the Predicted Quality channel assignment method to predict $SINR_{k,i}$, the uplink signal-to-interference-plus-noise ratio experienced by each co-channel active subscriber i on each conventional channel k if the new connection was to be assigned to channel k, without actually assigning the call to any conventional channel. Denote the

uplink transmit signal power and uplink spatial signature of co-channel active subscriber i on conventional channel k by $r_{k,i}$ and $a_{k,i}$, respectively. Then for each co-channel active subscriber i on each conventional channel k, the predicted spatial demultiplexing weights $\tilde{w}^U_{k,i}$, the predicted uplink signal power $S^U_{k,i}$ and the predicted uplink interference-plus-noise power $I^U_{k,i}$ are computed as

$$\tilde{w}^U_{k,i} = (\tilde{R}^{(k),i}_{zz})^{-1} r_{k,i} a_{k,i} \qquad \textbf{Eq. 11}$$

$$S^U_{k,i} = \left|\tilde{w}^{U\,*}_{k,i} a_{k,i}\right|^2 r_{k,i} \qquad \textbf{Eq. 12}$$

$$I^U_{k,i} = \left|\tilde{w}^{U\,*}_{k,i} \tilde{R}^{(k),i}_{zz} \tilde{w}^U_{k,i} - S^U_{k,i}\right| \qquad \textbf{Eq. 13}$$

where $\tilde{R}^{(k),i}_{zz}$ denotes the predicted sample covariance matrix at the base station currently communicating with active subscriber i on conventional channel k if the new subscriber was to be assigned to, and become active on the channel. $\tilde{R}^{(k)}_{zz}$ is computed similar to Eq. (7):

$$(\tilde{R}^{(k),i}_{zz})^{-1} = (\hat{R}^{(k),i}_{zz})^{-1} - (\hat{R}^{(k),i}_{zz})^{-1} a^i_k a^{i*}_k (\hat{R}^{(k),i}_{zz})^{-1} / ((1/r^{(k)}_{ss}) + a^{i*}_k (\hat{R}^{(k),i}_{zz})^{-1} a^i_k)$$

where a^i_k is the uplink spatial signature of the new subscriber to the base station currently communicating with active subscriber i on conventional channel k and $\hat{R}^{(k),i}_{zz}$ is the sample covariance matrix at the base station currently communicating with active subscriber i on conventional channel k. The predicted uplink SINR of active subscriber i on channel k, $SINR_{k,i}$, is then computed as $SINR_{k,i} = S^U_{k,i} / I^U_{k,i}$.

The Predicted Quality channel assignment method may optionally then be further constrained to only permit assignment of the new subscriber to conventional channel k if the predicted uplink SINRs of all active subscribers on that channel exceed a prescribed threshold level.

The Hierarchical method combines the advantages of the Weighted Correlation method and the Predicted Quality method for conventional channel assignment by applying the low complexity Weighted Correlation method as a means

for selecting a small subset of the least costly conventional channels and then applying the more optimal Predictive Quality method to the small subset for selecting the best conventional channel.

In the preceding description, consideration has been given to the assignment of uplink channels from the subscribers to the base station. Now consideration is given to the assignment of downlink channels from the base station to the subscriber stations.

During the initial phase of a call setup , for example, the subscriber station can measure the downlink received power levels on all of the conventional channels and report the measurements to the base station. This method is preferred if it does not introduce excessive latency (setup time). Alternatively, each subscriber station can periodically poll the downlink received power levels on all conventional channels whenever it is not actively making a call. The received power levels can be sent back to the base station over idle channels at prescribed intervals, or the power levels can be stored and updated at the subscriber station, and then all or a subset of the power levels communicated to the base station at the time of a call setup. Using these methods, the base station has a recent record of received power levels on all or a subset of the conventional channels for new subscribers.

In one embodiment, the received power level measured by the new subscriber on conventional channel k, P_k , is used as an estimate of I_k^D, the downlink interference-plus-noise power on conventional channel k. Alternatively, the estimate of I_k^D may be further refined for any conventional channel k already supporting one or more existing subscribers in the same cell as the new subscriber, by introducing a test interval. The duration of the test interval is typically chosen to be a small multiple (*e.g.*, between one and five) of the period between updates of the downlink multiplexing weights by the SDMAP in Fig. 1. During this test interval, the existing subscribers on channel k adjust their multiplexing weights as though the new subscriber has already been assigned to channel k. Methods for computing spatial multiplexing weights are described in Barratt *et al.*, U.S. Patent Application No. 08/375,848. In the latter part of the test interval, the new subscriber station may measure the downlink received power

on channel k and report the measurement to the base station. At the conclusion of the test interval, the existing subscribers on channel k readjust their multiplexing weights as though the new subscriber has been removed from channel k. The base station may then use P_k', the downlink received power measured during the test interval, as a refined estimate of I_k^D.

The cost function for conventional channel k is defined to be I_k^D. In one embodiment, the new subscriber's call is assigned to a conventional channel with an acceptable cost. In another embodiment, the new subscriber's call is assigned to the conventional channel for which the cost function is minimal. The assignment of more than one subscriber to the same conventional channel is permitted if there are sufficient hardware resources at the selected channel to accommodate the new connection in additional to any existing connections.

A further improvement in downlink channel assignment is obtained by predicting the quality of communication that will result from assigning a new connection to a particular conventional channel. This is accomplished by predicting the downlink SINR that a new subscriber will experience on each conventional channel if assigned to that channel, without necessarily assigning the call to any conventional channel.

The spatial signature of a new subscriber on the downlink is estimated for each conventional channel. In the case of a time-division-duplex (TDD) system, the downlink spatial signature of a new subscriber can be related to the uplink signature through calibration of the two links. A description of the calibration method is found in Roy *et al.*, U.S. Pat. No. 5,546,090. Other methods for estimating downlink spatial signatures are found in Barratt *et al.*, U.S. Pat. Application No. 08/375,848.

The downlink spatial multiplexing weights, w_k^D, for conventional channel k are then estimated. In a TDD system, the downlink weights w_k^D can be related to the previously described uplink spatial multiplexing weights, w_k^U, of Eq. (6) through calibration of the two links (Roy *et al.*, U.S. Pat. No. 5,546,090). Other methods for

estimating downlink weights are found in Barratt *et al.*, U.S. Pat. Application No. 08/375,848.

Having obtained for conventional channel k the downlink spatial signature, a_k^D, and the spatial multiplexing weight vector, w_k^D, the downlink signal power received by the new subscriber on conventional channel k, S_k^D, can be predicted as

$$S_k^D = \left| w_k^{D*} a_k^D \right|^2 \qquad \textbf{Eq. 14}$$

The downlink interference-plus-noise power level I_k^D may be estimated by P_k or P_k', as described above. The choice of P_k has the advantage of being minimally disruptive to existing subscribers, whereas P_k' offers greater accuracy. An alternative method for estimating I_k^D combines the advantages of P_k and P_k', at the expense of more computations but without using a test interval. The method starts with the assumption that P_k is known, and constructs a model for P_k as:

$$P_k = N_k + \sum_j \left\| W_{k,j}^{D*} a_k^{D,j} \right\|^2 \qquad \textbf{Eq. 15}$$

where $a_k^{D,j}$ is the downlink spatial signature on conventional channel k from base station j to the new subscriber whose multiplexing weights are unknown, N_k is the contribution to received power from noise and interference, $W_{k,j}^D$ is the multiplexing weight matrix formed by column-wise concatenating the multiplexing weights of each active subscriber (expressed as a column vector) on conventional channel k and served by base station j, and the summation is computed over all base stations j for which the weight matrices $\{W_{k,j}^D\}$ and spatial signatures $\{a_k^{D,j}\}$ are known. For example, $\{W_{k,i}^D\}$ may consist of the weights of the active subscribers on channel k served by the same base station as that of the new call, and $\{a_{k,i}^D\}$ the corresponding transmit spatial signature on channel k from this base station to the new call. $W_{k,j}^D$, the multiplexing weight matrix for active subscribers on channel k served by base station j which account for the presence

of the new subscriber, are then computed. There are many ways $W_{k,j}^{D}$ can be computed. For example, $W_{k,j}^{D}$ may be formed as

$$W_{k,j}^{D^*} = S_{k,j} A_{k,j}^{\dagger} \qquad \textbf{Eq. 16}$$

where $A_{k,j}$ is a matrix formed by column-wise concatenation of the known spatial signatures at base station j of the active subscriber on conventional channel k, $A_{k,j}^{\dagger}$ is the pseudoinverse of $A_{k,j}$ (see "Matrix Computations", Golub *et al.*, The Johns Hopkins University Press, Baltimore, MD, 1983), $S_{k,j}$ is a diagonal matrix of the signal amplitudes (which may be zero if the subscriber is not communicating with base station *j*). Note that U.S. Patent Application No. 08/375,848 uses matrix and vector notations which differ from the notation used in the present invention. However, such differences will be clear from the context to those of ordinary skill in the art.

In the illustrative embodiment, $\tilde{W}_{k,j}^{D}$ may then be computed from

$$\begin{bmatrix} \tilde{W}_{k,j}^{D^*} \\ w^* \end{bmatrix} = \tilde{S}_{k,j} \begin{bmatrix} A_{k,j} & a_k^{D,j} \end{bmatrix}^{\dagger} \qquad \textbf{Eq. 17}$$

where $a_k^{D,j}$ is the downlink spatial signature on conventional channel k from base station j to the new subscriber, $\tilde{S}_{k,j}$ is a diagonal matrix of transmit signal amplitudes, and $\tilde{W}_{k,j}^{D^*}$ is the submatrix formed by excluding the bottom row of $\begin{bmatrix} \tilde{W}_{k,j}^{D^*} \\ w^* \end{bmatrix}$. The downlink interference-plus-noise power I_k^D may then be predicted as

$$I_k^D = P_k - \sum_j \left\| W_{k,j}^{D^*} a_k^{D,j} \right\|^2 + \sum_j \left\| \tilde{W}_{k,j}^{D^*} a_k^{D,j} \right\|^2 \qquad \textbf{Eq. 18}$$

Having obtained I_k^D by any of the three methods described above, and having computed S_k^D from Eq. (14) for all conventional channels k, the predicted downlink signal-to-interference-noise-ratio for channel k (SINR) for each channel k is computed as

$$SINR_k^D = S_k^D / I_k^D \qquad \textbf{Eq. 19}$$

In one embodiment, the cost function for conventional channel k is computed as I_k^D. The new subscriber is assigned to the first conventional channel for which the computed cost function is acceptably low, or else to the conventional channel with the minimal cost. In an alternative embodiment, the cost function for any conventional channel k is computed as $-SINR_k^D$, the negative of the predicted SINR on that channel. The new subscriber is assigned either to the first conventional channel for which the computed cost function is acceptably low, or else to the conventional channel with the minimal cost.

The embodiments described in the preceding paragraph may be further constrained to only consider a particular conventional channel k as a candidate for assignment if one or more of the following conditions hold:

(1) $SINR_k^D$ is greater than a prescribed threshold, typically set at or near the SINR required to maintain an acceptable bit-error-rate for connection between the subscriber station and the base-station;

(2) the total transmit power on conventional channel k, including all active spatial channels, does not exceed the usable dynamic range of the base station RF power amplifier; and

(3) the predicted downlink signal to interference and noise ratio, $SINR_{k,i}^D$ experienced by each active subscriber, i, on conventional channel k, if the new connection were to be assigned to conventional channel k, is greater than some prescribed threshold.

The predicted downlink signal to interference and noise ratio for active subscriber i on conventional channel k, $SINR_{k,i}^D$ may be computed as follows. The downlink received power for active subscriber i on conventional channel k is denoted by $P_{k,i}$ and can be modeled by

$$P_{k,i} = N_{k,i} + \sum_j \left\| W_{k,j}^{D*} a_{k,i}^{D,j} \right\|^2 \qquad \textbf{Eq. 20}$$

where $a_{k,i}^{D,j}$ is the downlink spatial signature on conventional channel k from base station j to active subscriber i, $N_{k,i}$ is the unmodeled contribution to received power from noise and interferers for active subscriber i on conventional channel k. Then for each base station j the predicted spatial multiplexing weight matrix $\tilde{W}_{k,j}^{D}$ for conventional channel k that accounts for the presence of the new subscriber is computed. The column of the multiplexing weight matrix corresponding to active subscriber i on conventional channel k is denoted $\tilde{w}_{k,j}^{D}$. The predicted downlink signal power $S_{k,i}^{D}$ and the predicted downlink interference-plus-noise power $I_{k,i}^{D}$ are computed as

$$S_{k,i}^{D} = \left| \tilde{w}_{k,i}^{D*} a_{k,i}^{D} \right|^2 \qquad \textbf{Eq. 21}$$

$$\begin{aligned} I_{k,i}^{D} &= N_{k,i} + \sum_j \left\| \tilde{W}_{k,j}^{D*} a_{k,i}^{D,j} \right\|^2 - S_{k,i}^{D} \\ &= P_{k,i} - \sum_j \left\| W_{k,j}^{D*} a_{k,i}^{D,j} \right\|^2 + \sum_j \left\| \tilde{W}_{k,j}^{D*} a_{k,i}^{D,j} \right\|^2 - S_{k,i}^{D} \end{aligned} \qquad \textbf{Eq. 22}$$

The predicted downlink SINR of active subscriber i on conventional channel k, $SINR_{k,i}^{D}$ is then computed as $SINR_{k,i}^{D} = S_{k,i}^{D} / I_{k,i}^{D}$.

As previously mentioned, the effect of intermodulation distortion is an important consideration for wideband radio transmitters and receivers, and can be taken into consideration during channel assignment by two methods based on the crest factor of the composite RF signal:

(1) augmenting the cost function for penalizing channel selections that are likely to make significant increases in intermodulation distortion as predicted by a significant increase in the crest factor; or

(2) adding a constraint in the channel assignment process that prohibits the selection of any channel that causes the crest factor of the composite wideband RF signal to exceed a prescribed acceptable level.

For sake of clarity, in the ensuing discussion a wideband radio transmitter will be associated with the downlink channel and a wideband receiver with the uplink channel. Those of ordinary skill in the art will recognize that the channel assignment methods described herein which account for nonlinear distortion can be applied to a wideband transmitter or receiver employed on either uplink or downlink. In a wideband radio transmitter, intermodulation distortion levels are most severe when one or a few downlink channel carrier frequencies (sub-bands) are transmitting at much higher levels than the rest, because the intermodulation products tend to have prominent spectral peaks while composite wideband RF signals having more uniform spectral distributions tend to produce more uniform broadband intermodulation products. An analogous situation holds at a wideband radio receiver, wherein intermodulation distortion produced by RF mixers, low noise amplifiers, etc. tend to be the most severe if one or a few uplink channel carrier frequencies (sub-bands) have much higher received power levels than the rest. The crest factor, which is defined as a ratio of peak power in a given sub-band of a broadband signal to the power averaged over the total bandwidth of the composite RF signal, is a measure of magnitude of the offending spectral peaks. Because the assignment of spatial channels can significantly increase the transmit (received) power on a given downlink (uplink) conventional channel subcarrier by assigning multiple subscribers with different spatial signatures to the same conventional channel, the crest factor is an important tool for predicting intermodulation distortion. Also, a wideband radio system which uses TDMA can create large temporal power peaks during some time slots due to the subscriber channel assignments. Both spectral and temporal cresting are important considerations.

For a combined TDMA/FDMA wideband radio, control of intermodulation distortion due to temporal and frequency cresting is based on predicting the crest factor that would result by assigning a subscriber to a particular downlink or uplink conventional channel, time slot, and spatial channel. The uplink (downlink) temporal crest factor, C_i, during uplink (downlink) time slot i is defined as

$$C_i = \frac{\max_{l=1,\ldots,L}\{\sum_{j=1}^{d_{i,l}} P_{i,j,l}\}}{(1/L)\sum_{l=1}^{L}\sum_{j=1}^{d_{i,l}} P_{i,j,l}} \qquad \textbf{Eq. 23}$$

where L is the number of uplink (downlink) carrier frequencies, $d_{i,l}$ is the number of spatial channels on uplink (downlink) carrier frequency l and on uplink (downlink) time slot i, and $P_{i,j,l}$ is the receive (transmit) power of a subscriber on uplink (downlink) carrier frequency l, uplink (downlink) time slot i, and uplink (downlink) spatial channel j. Let the maximal uplink (downlink) crest factor over uplink (downlink) timeslots i be:

$$C_{\max} = \max_{i} C_i \qquad \textbf{Eq. 24}$$

In the preferred embodiment, the maximal uplink (downlink) crest factor, $C_{\max}$, may be used by any of the previously described uplink (downlink) channel assignment methods as an additional constraint, by comparing its value with a prescribed threshold level, and prohibiting the channel selection if the threshold level is exceeded. The prescribed threshold level is obtained by measuring the intermodulation distortion level of the particular MCPA as a function of crest factor, in the case of the downlink; or by measuring the intermodulation distortion level of the aggregate analog receiver chain, in the case of the uplink.

Alternatively, the cost function c_k for conventional channel k as computed by any previously described channel assignment method may be augmented by the crest factor as follows:

$$c_k' = c_k + \lambda C_{\max} \qquad \textbf{Eq. 25}$$

where λ is a user-defined constant that determines the relative importance of the crest factor $C_{\max}$ with respect to c_k.

While the description for the consideration of intermodulation distortion in channel assignment has been in terms of a combined TDMA/FDMA system, the invention is equally applicable to pure TDMA or FDMA systems, which may be treated as special cases of the combined system.

Fig. 3(a) is a flow diagram of the Weighted Correlation method 300 for channel assignment as previously described. In step 301, the channel index k is initialized. The hardware resources for conventional channel k are checked in step 302,

and, if not adequate, the channel index is incremented in step 309 and the process returns to step 302. If the hardware resources are adequate, the receive spatial signature of the new subscriber on candidate channel k is estimated or else obtained from a database in step 303. The covariance matrix is computed in step 304 using the methods of Figs. 3(b) or 3(c). The cost function for conventional channel k is computed in step 305 in accordance with Eq. (1), using the spatial signature of the new subscriber from step 303 and the covariance matrix from step 304. In step 306, conventional channel k is checked to see if any additional constraints (*e.g.*, the computed cost is below a prescribed threshold, and/or the uplink maximal crest factor is less than a prescribed threshold) are satisfied and, if not, the process goes to step 308. Otherwise, conventional channel k is added to a candidate list of channels to be considered for assignment in step 307. If, in step 308, all channels have been examined, the process moves to step 310. Otherwise, the process returns to step 309 to increment the channel index and for another iteration through steps 302-308. Step 310 checks if the candidate list has any candidate channels and, if not, the call is not assigned at this base station in step 311, that is, no assignment is made. Otherwise, a best channel k is selected in step 312. In one embodiment, the best channel is any channel with a cost that is less than a prescribed minimum, *e.g.*, the first channel found to have a cost less than the prescribed minimum. (For example, the prescribed minimum can be chosen as a cost level that corresponds to a maximum allowable bit error rate between a subscriber and base station.) In the preferred embodiment, the best channel is the minimal cost channel. If, in step 313 , it is determined that selected channel k is not in use within the cell, conventional channel k is assigned to the new subscriber in step 314. If conventional channel k is in use by a subscriber within the cell, a spatial channel, using conventional channel k, is assigned to the new subscriber in step 315.

Fig. 3(b) is a flow diagram for the unstructured method 370 for estimating the sample covariance matrix $\hat{R}_{zz}^{(k)}$. In step 371, measurements are made of the base station received signal vectors, $\{z^{(k)}(i)\}$, from the antenna array for each conventional channel k, and in step 372 the estimate, $\hat{R}_{zz}^{(k)}$, is computed using Eq. (2).

Fig. 3(c) is a flow diagram of the structured method 350 for estimating the sample covariance matrix $\tilde{R}_{zz}^{(k)}$. In step 351, the transmit (TX) powers of the active co-channel subscribers for each conventional channel are retrieved from a database or else measured. In step 352, the spatial signatures of co-channel active subscribers are estimated. In step 353, the quantity $A_k^* R_{ss}^{(k)} A_k$ is computed where A_k and $R_{ss}^{(k)}$ are as defined above with respect to Eq. (3). In step 354, the noise-plus-interference covariance matrix, $R_{nn}^{(k)}$, may be estimated by measuring the received noise and intercell interference signal at each antenna element receiver or by using the spatial signatures and transmit powers of all active subscribers outside of the base station cell, if available, as described previously with respect to Eq. (4). The structured estimate of the sample covariance matrix $\tilde{R}_{zz}^{(k)}$ is then computed in step 355 by using Eq. (3).

Fig. 4 is a flow diagram of Predictive Channel Assignment Method 400 in which it is assumed that the covariance matrix for each conventional channel k, $\hat{R}_{zz}^{(k)}$, its inverse, $(\hat{R}_{zz}^{(k)})^{-1}$, the new subscriber transmit power, r_{ss}, and the new subscriber spatial signature, a_k, are known. The channel index is initialized (k=1) in step 401. In step 402, channel k is checked for the required hardware resources and if not adequate, the channel index is incremented in step 411 and the process returns to step 402. Otherwise, the existing covariance matrix, $\hat{R}_{zz}^{(k)}$, is updated to $\tilde{R}_{zz}^{(k)}$ in step 403 by including the predicted effects of the presence of the new subscriber in accordance with Eq. (5). In step 404, the inverse matrix $(\tilde{R}_{zz}^{(k)})^{-1}$ is computed in accordance with Eq. (7) and is then used in step 405 to compute the SDMA demultiplexing weights, w_k^U, in accordance with Eq. (6). In step 406, the predicted received uplink power, S_k^U, interference, I_k^U, and SINR that would result if the new connection were to be assigned to conventional channel k are computed for all conventional channels using Eqs. (8), (9) and (10), respectively. In step 407, a cost is computed (such as $c_k = I_k^U$ or $-SINR_k$). In step 408, conventional channel k is checked to see if any additional constraints are met. In one embodiment, the constraint includes the predicted SINR of the new connection being greater than a prescribed threshold. Additionally, a further constraint may include the predicted SINRs of active calls on conventional channel k, computed using Eqs.

(11), (12) and (13), also exceeding some prescribed threshold. An optional constraint of the uplink maximal crest factor being less than some prescribed threshold may also be imposed. If any of the constraints are not satisfied, the process goes to step 410. Otherwise, conventional channel k is added to the candidate list in step 409. In step 410, a check is made as to whether all conventional channels have been processed and, if not, the channel index is incremented in step 411 and the process iterates through steps 402-410. Otherwise, a check is made in step 412 to determine if the candidate list is empty and, if so, the new connection at this base station is left unassigned (step 413). If the candidate list is not empty, a selection is made in step 414 of the channel satisfying the prescribed requirement for best channel (as previously discussed). If, in step 415, it is found that selected channel k is in use within the cell, a spatial channel using conventional channel k is assigned to the new connection in step 417. Otherwise, conventional channel k is assigned to the new connection in step 416.

Fig. 5 is a flow diagram for Hierarchical Method 500 for uplink channel assignment. In step 501, the Weighted Correlation Assignment Method 300, steps 301-309 are invoked for producing a set of candidate channels together with their costs. In step 502, a subset of low cost conventional channels is selected from the set of all candidates. In step 503, Predictive Assignment Method 400 is applied to the subset of conventional channels provided by step 502 for selecting the best conventional channel for assignment.

Fig. 6 is a flow diagram of Downlink Channel Assignment Method 600, as described previously. The channel index, k, is initialized in step 601. In step 602, a determination is made as to whether or not channel k has the required hardware resources, and, if not, the process increments the channel index in step 603 and then returns to step 602 for another iteration. Otherwise, in step 604, the new subscriber reports received powers level for each conventional channel k to the base station. The measurement of the received power may optionally be performed during a test interval, as previously described. The base station, in step 605, determines the downlink interference-plus-noise power level, I_k^D, by any of the three methods described earlier. In step 606, an option is exercised: if downlink channel assignment is to be based on predicting SINR levels, the process moves to step 609; otherwise, the process moves to

step 611. The base station, in step 609 estimates the downlink received power level, S_k^D, from Eq. (14) using the spatial signature a_k^D and multiplexing weights w_k^D of the new subscriber on conventional channel k. The cost is computed in step 610 as $c_k = -S_k^D / I_k^D$, *i.e.*, the negative of $SINR_k^D$ defined in Eq. (19). If the SINR option is not selected in step 606, a cost based on interference-plus-noise is computed, *i.e.*, $c_k = I_k^D$. In step 612, a determination is made as to whether all constraints are satisfied by channel k. In one embodiment, the constraint includes the predicted SINR of the new connection being greater than a prescribed threshold. Additionally, a further constraint may include the predicted SINRs of active calls on conventional channel k also exceeding some prescribed threshold. An optional constraint may also be imposed of the downlink maximal crest factor being less than some prescribed threshold. See the above equations for calculation of $SINR_{k,i}^D$ for details for the different alternatives for the determination of step 612. If in step 612, any of the constraints are not satisfied, the process goes to step 614. Otherwise, the process goes to step 613 where channel k is added to the candidate channel list. If, in step 614, it is determined that all channels have not been considered, the process goes to step 603. Otherwise, a check is made in step 615 to determine if the candidate list is empty and, if so, the new call is left unassigned at this base station (step 616). If the candidate list is not empty, a selection is made in step 617 of the channel satisfying the prescribed requirement for best channel (as previously discussed). If, in step 618, it is found that selected channel k is in use within the cell, a spatial channel using conventional channel k is assigned to the new subscriber in step 620. Otherwise, conventional channel k is assigned to the new subscriber in step 619.

Because many practical systems impose a fixed relationship between the uplink and downlink of a full-duplex conventional channel assignment, it is necessary to define a method for joint uplink-downlink channel assignment. This is accomplished by selecting a subset of those uplink and downlink pairs that satisfy both the uplink assignment constraints (Figs. 3(a-c), 4, and 5, and as previously described) and the downlink assignment constraints (Fig. 6 and as previously described). For example, this may mean selecting those full-duplex channels with estimated uplink and downlink

SINRs that are above the prescribed thresholds and that satisfy the crest factor constraint. From this subset of uplink/downlink pairs, an uplink cost, c_k^U, and a downlink cost, c_k^D, is computed for each pair and then combined to form a single uplink/downlink cost. As is well known to those skilled in the art, there are many possible ways to combine the individual uplink or downlink costs of a given pair of uplink and downlink costs. For example, the weighted sum $(c_k^U + \lambda c_k^D)$, where λ is a relative scaling factor, can be used for the joint cost, or the weighted product $c_k^U (c_k^D)^\gamma$, where γ is a relative exponential weighting factor, can be used. A reasonable joint cost function, the uniformly weighted $(\gamma = 1)$ joint product cost function is

$$c = c_k^U \cdot c_k^D \qquad \textbf{Eq. 26}$$

(Because of the many possible choices available for combining the uplink and downlink costs, it is noted that the concept of creating a joint cost is important rather than the specific form of the joint cost function selected.) Having created a joint cost for each full-duplex channel pair in the selected subset, the full-duplex channel with either an acceptably low joint cost or the minimum joint cost is selected for assignment.

All of the channel assignment methods previously described may be applied to either a set of candidate conventional channels associated with any particular base station or else a set of candidate conventional channels associated with a multiplicity of base stations. In the latter embodiment, the channel assignment method automatically performs the selection of a base station for the new connection in the process of selecting a conventional channel for the call. This can be achieved either by assigning the channel and associated base station that provide the best cost from all candidate channels at all associated base stations or sequentially carrying out the channel assignment at candidate channels until a conventional channel is assigned.

Channel reassignment may be necessary if a newly admitted call experiences communication quality problems, or if an in-progress call experiences an unacceptable reduction in quality due to a change in its RF environment. The channel reassignment process is the same as for initial channel assignment, except that the conventional channel the subscriber wishes to be reassigned from is removed from the

list of candidate channels prior to the selection process. With this modification, all of the methods thus described for channel assignment are also applicable to channel re-assignment.

Call admission control is the decision process for admitting or not admitting a new connection. It may be necessary to not assign a new connection if the system load is very high such that admitting a new connection may have a significant negative impact on the quality of the existing connections in the system. For any of the previously described channel assignment methods, the determination as to whether the assignment of a new connection to any particular channel will have a significant negative impact on existing connections can be done by checking the constraints of the particular channel assignment method. If the constraints can be met for the new connection on conventional channel k, then this channel is a candidate for assignment. Otherwise, channel k is not permissible for assignment. Call admission control is accomplished by checking the constraints for all or a subset of conventional channels, and if no channel satisfying the constraints of the particular channel assignment method can be found, then not assigning the new connection. Hence, all of the previously described channel assignment methods are also applicable for admission control. Additionally, an alternate method which can be applied independently of any channel selection method may also be used. Fig. 7, a flow diagram for Call Admission Control Method 700, summarizes the different methods.

The first method, appropriate when the Weighted Correlation channel assignment is used, begins at step 701 where a cost threshold value is established which is compared with the expected cost, c_k, of a selected channel k, in step 702. If the cost threshold is not exceeded, channel k is assigned in step 703. If the cost threshold is exceeded, the assignment of channel k is left unassigned in step 704.

The second method, appropriate when the Uplink Predicted Quality or Downlink Predictive method for channel assignment is used, begins at step 721 (Fig. 7) where a SINR threshold level is assigned. In step 722, the SINR of the conventional channel selected for assignment is compared with the threshold and, if less than the

threshold, the selected channel is assigned in step 723. Otherwise, the selected channel assignment is left unassigned in step 724.

The third method, appropriate for any assignment method and SDMA and non-SDMA systems, begins at step 711 of Fig. 7 where a load threshold is assigned. In step 712, the cell system load is measured and then, in step 713, compared with the load threshold. If load threshold is not exceeded, the selected channel is assigned in step 714. If the load threshold is exceeded, all conventional channels are left unassigned in step 715 until the system load falls below the load threshold value. A method for measuring system load is by monitoring the rate of intercell handoffs, or the rate of channel reassignments (intracell handoffs), that is experienced by the cell. A moving time average of the rate of handoffs to other cells, or the rate of channel reassignments within the cell, can be used for smoothing the stochastic behavior of these events.

Fig. 7 describes the assignment locally, that is, at one base station. Recalling that overall assignment may be carried out either sequentially or by "joint optimization," when the assignment is carried out sequentially among *all* candidate conventional channels at *all* associated base stations, steps 704, 715, and 724 would each be followed by a channel assignment process at the next base station, and such a process would be repeated at different base stations until a conventional channel and associated base station are found, or the call is left unassigned. When the assignment is carried out by joint optimization, that is, by assigning the channel and associated base station that provide the best cost from *all* candidate channels at *all* associated base stations, then the assignment steps in Fig. 7 would be modified to carry out assignment of both the conventional channel and the associated base station, and how to carry out such modification to the flow diagram of Fig. 7 would be clear to those of ordinary skill in the art.

The methods described above were, for sake of clarity in the description, limited to specific wireless cellular communication systems and embodiments but, for those of ordinary skill in the art, the application of these inventions to other similar communication systems, such as wireless local area networks, and to other variations on the embodiment will become evident to those practicing the art from the descriptions

provided without departing from the spirit and scope of the invention which should only be limited as set forward in the claims that follow.

CLAIMS

What is claimed is:

1. A channel assignment method for use in a wireless communication system for establishing an uplink connection between a new subscriber station requesting an uplink connection and a base station, and for a base station reassigning an existing connection to a subscriber station, the base station assigning a conventional channel to the subscriber station based on a cost function that is indicative of the uplink interference that may be realized by assigning a conventional channel, and for controlling interference from existing connections by selecting a conventional channel with an acceptably low cost, the method comprising:

   (a) creating a cost function that is indicative of the interference level expected from existing connections on each conventional channel;

   (b) computing a cost for each conventional channel by using the cost function; and

   (c) assigning a conventional channel to the subscriber station by selecting a conventional channel with a cost that is less than a prescribed cost threshold.

2. The method of claim 1 further comprising the step of assigning a spatial channel to the assigned conventional channel if the assigned conventional channel is in use.

3. The method of claim 1 wherein the assigned conventional channel has a minimal cost.

4. The method of claim 1 wherein the cost function is a quadratic function of the uplink spatial signature of the subscriber station.

5. The method of claim 4 wherein the cost function is a weighted quadratic function of the uplink spatial signature of the subscriber station.

6. The method of claim 5 wherein the cost function is weighted by a sample covariance matrix of the base station antenna array received vector signal for a conventional channel.

7. The method of claim 1 wherein the cost function, c_k, is computed as

$$c_k = a_k^* \hat{R}_{zz}^{(k)} a_k$$

where a_k is the conventional channel k uplink spatial signature of the subscriber station, and $\hat{R}_{zz}^{(k)}$ is a sample covariance matrix of the base station antenna array response on channel k.

8. The method of claim 7 wherein the sample covariance matrix, $\hat{R}_{zz}^{(k)}$, is estimated as an average from the received signal vectors, $z^{(k)}(i)$, at the base station antenna array on conventional channel k, where i is a received signal vector sample index, as

$$\hat{R}_{zz}^{(k)} = \frac{1}{n}\sum_{i=1}^{n} z^{(k)}(i) z^{(k)*}(i) .$$

9. The method of claim 7 wherein the sample covariance matrix, $\hat{R}_{zz}^{(k)}$, is estimated as

$$\hat{R}_{zz}^{(k)} = A_k R_{ss}^{(k)} A_k^* + R_{nn}$$

where A_k is a collection of spatial signatures formed by column-wise concatenation of all spatial signatures of subscriber stations actively communicating with the base station on conventional channel k, $R_{ss}^{(k)}$ is a covariance matrix whose diagonal elements are average transmit powers of subscribers communicating with the base station, and $R_{nn}^{(k)}$ is a noise-plus-interference covariance matrix of the base station antenna array received signals.

10. The method of claim 9 wherein $R_{nn}^{(k)}$ is estimated as

$$R_{nn}^{(k)} = A_k^0 R_{s_0 s_0}^{(k)} A_k^{0*} + \sigma^2 I$$

where A_k^0 is a collection of spatial signatures of subscriber stations not communicating with the base station and is formed by column-wise concatenation of the spatial signatures on conventional channel k, and $R_{s_0 s_0}^{(k)}$ is a covariance matrix whose diagonal elements are the average transmit signal powers of subscriber stations not communicating with the base station, σ^2 is an estimated receiver noise power, and I is an identity matrix.

11. The method of claim 1 wherein step (a) for creating a cost function is based on at least one uplink received signal quality factor selected from the group consisting of a received uplink signal level, a signal-to-interference-plus-noise ratio, an interference-plus-noise level, an intermodulation noise level, or a crest factor.

12. A channel assignment method for use in a wireless communication system for establishing an uplink connection from a new subscriber station requesting an uplink connection to a base station, and for reassigning an existing connection to a subscriber station, by assigning a conventional uplink channel to the new subscriber station based on a system model for predicting expected uplink signal quality levels on each conventional channel from existing connections, and for controlling interference from existing connections by selecting a conventional channel with an acceptable predicted quality level, the method comprising:

(a) creating a model of the wireless communication system for predicting an uplink received signal quality level on each conventional uplink channel based upon existing connections;

(b) predicting the uplink received signal quality on each uplink conventional channel based upon the model and upon existing connections;

(c) computing a predicted cost for each uplink conventional channel, based on the predicted received signal quality using a prescribed cost function; and

(d) assigning an uplink conventional channel to the new subscriber station that has a predicted cost that is less than a prescribed cost level.

13. The method of claim 12 further comprising the step of assigning a spatial channel to the assigned conventional channel if the assigned conventional channel is in use, and if the wireless communication system supports more than one spatial channel per conventional channel.

14. The method of claim 12 wherein the predicted received signal quality is based on predicting an uplink interference-plus-noise level.

15. The method of claim 12 wherein the predicted received signal quality is based on predicting a received signal to interference-plus-noise ratio (SINR).

16. The method of claim 12 wherein the prescribed cost level corresponds to a minimal acceptable received bit error rate.

17. The method of claim 12 wherein the assigned uplink conventional channel has a predicted minimal cost level.

18. The method of claim 12 wherein step (d) further requires that the assigned conventional channel have a SINR level that is greater than a prescribed minimal value.

19. The method of claim 12 wherein step (a) for creating a model of the wireless communication system further comprises:

(i) updating a covariance matrix of received uplink signals, $\hat{R}_{zz}^{(k)}$, representative of received uplink signals at the base station before assigning channel k by predicting an updated covariance matrix, $\tilde{R}_{zz}^{(k)}$, representative of received uplink signals that would result if the new subscriber were to be assigned to channel k, where $\tilde{R}_{zz}^{(k)} = \hat{R}_{zz}^{(k)} + a_k r_{ss}^{(k)} a_k^*$, a_k is the uplink spatial signature of the new subscriber on conventional channel k, and $r_{ss}^{(k)}$ is a scalar representative of the new subscriber transmitted power on conventional channel k; and

(ii) computing an uplink spatial demultiplexing weight vector, $\tilde{w}_k^U$, where $\tilde{w}_k^U = (\tilde{R}_{zz}^{(k)})^{-1} a_k r_{ss}^{(k)}$.

20. The method of claim 19 wherein the updated inverse covariance matrix, $\left(\tilde{R}_{zz}^{(k)}\right)^{-1}$, is obtained from the expression

$$(\tilde{R}_{zz}^{(k)})^{-1} = (\hat{R}_{zz}^{(k)})^{-1} - (\hat{R}_{zz}^{(k)})^{-1} a_k a_k^* (\hat{R}_{zz}^{(k)})^{-1} / ((1/r_{ss}^{(k)}) + a_k^* (\hat{R}_{zz}^{(k)})^{-1} a_k).$$

21. The method of claim 12 wherein step (b) for predicting the uplink received signal quality level on channel k comprises:

(i) predicting uplink received signal power as $S_k^U = \left|\tilde{w}_k^{U*} a_k\right|^2 r_{ss}^{(k)}$;

(ii) predicting uplink interference-plus-noise power for the new subscriber as $I_k^U = \tilde{w}_k^{U*} \hat{R}_{zz}^{(k)} \tilde{w}_k^U$; and

(iii) predicting the uplink SINR for channel k as $SINR_k = S_k^U / I_k^U$.

22. The method of claim 12 further comprising a constraint for only permitting assignment of a conventional channel if the assignment results in a predicted uplink SINR for all active subscribers using the conventional channel which exceeds a prescribed SINR level.

23. The method of claim 12 wherein the predicted cost is based on at least one set of predicted uplink received signal quality factors selected from the group consisting of received uplink signal levels, interference-plus-noise levels, intermodulation noise levels, and crest factor values.

24. A channel assignment method for use in a wireless communication system, for establishing an uplink connection from a new subscriber station requesting an uplink connection to a base station, and for reassigning an existing connection to a subscriber station, by assigning a conventional channel to the subscriber station based on using a cost function that is indicative of the interference that may be experienced on each uplink conventional channel for computing a cost for assigning any conventional channel, and for controlling interference from existing connections by selecting a subset of conventional channels with an acceptably low cost, and also by assigning a conventional channel to the subscriber station from the selected subset by predicting expected uplink interference levels on each conventional channel from existing connections by use of a system model, and for controlling interference from existing connections by selecting a conventional channel with an acceptably low interference level, the method comprising:

(a) creating a first cost function that represents the interference expected from existing connections on each conventional uplink channel;

(b) computing a first cost for each conventional uplink channel;

(c) selecting a subset of conventional uplink channels by selecting all conventional uplink channels with a first cost that is less than a prescribed first cost level;

(d) creating a model of the wireless communication system for predicting an uplink received signal quality level on each conventional uplink channel based upon existing connections;

(e) creating a second cost function representative of the predicted uplink received signal quality and based upon the model and upon existing connections;

(f) predicting the uplink received signal quality level on each conventional uplink channel of the subset based upon the second cost function;

(g) computing a second cost using the second cost function for each conventional uplink channel of the subset; and

(h) selecting a conventional uplink channel that has a predicted second cost that is less than a prescribed second cost level for assignment to the new subscriber station.

25. The method of claim 24 further comprising the step of assigning a spatial channel to the assigned conventional uplink channel if the assigned conventional uplink channel is in use, and if the wireless communication system supports more than one spatial channel per conventional channel.

26. The method of claim 24 wherein the prescribed first and second cost levels correspond to a prescribed maximal received uplink bit error rate level.

27. The method of claim 24 wherein step (h) further requires that the assigned conventional channel have a SINR level that is higher than a prescribed SINR level.

28. The method of claim 24 further providing that if the wireless communication system uses power control when establishing an uplink connection, a conventional uplink channel is selected for assignment that has a SINR level greater than a prescribed SINR level.

29. The method of claim 24 wherein step (e) for creating a second cost function comprises creating a cost function based on at least one set of uplink received signal

quality factors selected from the following list: received uplink signal power levels, interference-plus-noise levels, intermodulation noise levels, and crest factor values.

30. A channel assignment method for use in a wireless communication system for establishing a downlink connection between a base station and a new subscriber station by assigning, to the base station, a downlink conventional channel based on achieving an acceptable cost level using a cost function based on a downlink noise-plus-interference level from existing connections, the method comprising:

(a) estimating, at the base station, the downlink received interference-plus-noise level that would result for each existing conventional channel if the new subscriber was assigned to a given downlink conventional channel;

(b) computing, at the base station, a cost for each existing downlink conventional channel using a prescribed cost function based on the estimated downlink received interference-plus-noise levels; and

(c) assigning, at the base station, a conventional downlink channel that has a computed cost that is less than a prescribed value.

31. The method of claim 30 further comprising the step of assigning a spatial channel to the assigned conventional downlink channel if the assigned conventional downlink channel is in use, and if the wireless communication system supports more than one spatial channel per conventional downlink channel.

32. The method of claim 30 wherein step (c) further comprises selecting a downlink channel that has a minimal computed cost.

33. The method of claim 30 wherein step (a) further comprises:

(i) measuring, at the subscriber station, downlink received signal levels on each downlink channel and reporting the downlink received signal levels to the base station;

(ii) estimating, at the base station, the downlink received interference-plus-noise levels from the reported downlink signal levels of step (i);

34. The method of claim 33 wherein step (i) further comprises each subscriber station, when not actively engaged in a call, periodically measuring the downlink

received signal level on each conventional channel and reporting the downlink received signal levels to the base station when necessary.

35. The method of claim 33 wherein the downlink received interference-plus-noise levels is estimated as being the reported downlink received signal levels.

36. The method of claim 30 wherein step (a) for estimating downlink received interference-plus-noise levels on each downlink channel comprises the following steps:

(i) adjusting, at the base station, each existing subscriber's downlink multiplexing weights as if the new subscriber was assigned to a given conventional channel;

(ii) measuring, at the new subscriber station, the downlink received signal level on the given channel after step (i) and reporting the downlink received signal level to the base station;

(iii) predicting, at the base station, a downlink interference-plus-noise level from the downlink received signal level of step (ii); and

(iv) readjusting, at the base station, each existing subscriber's downlink multiplexing weights as if the new subscriber was not assigned to the given conventional channel.

37. The method of claim 30 wherein step (a) for estimating a downlink received interference-plus-noise level, I_k, for each existing channel k comprises, modeling the downlink received interference-plus-noise level, I_k, as a sum of the noise contribution, N_k, and a predicted second interference signal level that would result if the new subscriber was to be assigned to channel k, estimating the noise contribution N_k as a signal level difference between a measured downlink received signal level, P_k, on channel k and a computed first interference signal level due to all base stations using channel k.

38. The method of claim 37 wherein the computed first interference signal level is computed as $\sum_j \left\| W_{k,j}^{D*} a_k^{D,j} \right\|^2$, where $W_{k,j}^{D*} = S_{k,j} A_{k,j}{}^{\dagger}$, $S_{k,j}$ is a diagonal matrix of

base station j transmitted signal amplitudes, $A_{k,j}$ is a column-wise concatenated matrix of known downlink spatial signatures at base station j for conventional channel k, $A_{k,j}^{\dagger}$ represents the pseudoinverse of $A_{k,j}$, and $a_k^{D,j}$ is a downlink spatial signature for conventional channel k from base station j.

39. The method of claim 37 wherein the predicted second interference signal level for channel k is computed as $\sum_j \left\| \tilde{W}_{k,j}^{D*} a_k^{D,j} \right\|^2$, where $\begin{bmatrix} \tilde{W}_{k,j}^{D*} \\ w^* \end{bmatrix} = \tilde{S}_{k,j}\left[A_{k,j} a_k^{D,j}\right]^{\dagger}$, $[\cdot]^{\dagger}$ represents a pseudoinverse operation, $\tilde{S}_{k,j}$ is a diagonal matrix of transmit signal amplitudes, $A_{k,j}$ is a column-wise concatenated matrix of known downlink spatial signatures at base station j, $a_k^{D,j}$ is a downlink spatial signature for conventional channel k from base station j for conventional channel k, and w^* is the bottom row of matrix $\tilde{S}_{k,j}\left[A_{k,j} a_k^{D,j}\right]^{\dagger}$.

40. The method of claim 30 further comprising a step for assigning a selected downlink conventional channel only if it is expected to result in an acceptable downlink received signal quality for each active subscriber using the selected conventional channel based upon at least one downlink quality factor selected from a group consisting of: predicted cost ; predicted total base station transmitter power; predicted intermodulation distortion level; predicted interference-plus-noise level; and predicted signal to interference-plus-noise ratio .

41. The method of claim 40 wherein the predicted intermodulation distortion level is obtained by computing a crest factor for each downlink conventional channel, the crest factor value being indicative of an intermodulation distortion level that would result if a given downlink conventional channel were to be assigned.

42. A channel assignment method for use in a wireless communication system for establishing a downlink connection between a base station and a new subscriber station by assigning, to the base station, a downlink conventional channel based on achieving an acceptable cost level using a cost function based on a downlink signal

to noise-plus-interference (SINR) level from existing connections, the method comprising:

(a) estimating, at the base station, the downlink received SINR level that would result for each existing conventional channel if the new subscriber were to be assigned to a given downlink conventional channel;

(b) computing, at the base station, a cost for each existing downlink conventional channel using a prescribed cost function based on the estimated downlink received SINR level; and

(c) assigning, at the base station, a downlink channel that has a computed cost that is less than a prescribed value.

43. The method of claim 42 further comprising the step of assigning a spatial channel to the assigned conventional downlink channel if the assigned conventional downlink channel is in use, and if the wireless communication system supports more than one spatial channel per conventional downlink channel.

44. The method of claim 42 wherein the prescribed cost function of step (b) is based on $SINR_k^D$, an estimated downlink signal to interference-plus-noise ratio for each channel k, where $SINR_k^D = S_k^D / I_k^D$, $S_k^D = \left|w_k^{D*} a_k^D\right|^2$, w_k^D is the downlink multiplexing weight for the new subscriber on channel k, a_k^D is the downlink spatial signature of the new subscriber on channel k, and I_k^D is an estimated interference-plus-noise level.

45. The method of claim 42 wherein step (c) further comprises selecting a downlink channel that has a minimal computed cost.

46. The method of claim 42 wherein step (a) further comprises:

(i) measuring, at the subscriber station, downlink received signal levels on each downlink channel and reporting the downlink received signal levels to the base station;

(ii) estimating, at the base station, the downlink received interference-plus-noise level from the reported downlink signal levels of step (i).

47. The method of claim 46 wherein step (i) further comprises each subscriber station, when not actively engaged in a call, periodically measuring the downlink received signal level on each conventional channel and reporting the downlink received signal levels to the base station when necessary.

48. The method of claim 42 wherein step (a) for estimating downlink received interference-plus-noise levels on each downlink channel comprises the following steps:

(i) adjusting, at the base station, each existing subscriber's downlink multiplexing weights as if the new subscriber was assigned to a given conventional channel;

(ii) measuring, at the new subscriber station, the downlink received signal level on the given channel after step (i) and reporting the downlink received signal level to the base station;

(iii) predicting, at the base station, a downlink interference-plus-noise level from the downlink received signal level of step (ii); and

(iv) readjusting, at the base station, each existing subscriber's downlink multiplexing weights as if the new subscriber were not assigned to the given conventional channel.

49. The method of claim 42 wherein step (a) for estimating a downlink received interference-plus-noise level, I_k, for each existing channel k comprises, modeling the downlink received interference-plus-noise level, I_k, as a sum of the noise contribution, N_k, and a predicted second interference signal level that would result if the new subscriber was to be assigned to channel k, estimating the noise contribution N_k as a signal level difference between a measured downlink received signal level, P_k, on channel k and a computed first interference signal level due to all base stations using channel k.

50. The method of claim 49 wherein the computed first interference signal level is computed as $\sum_j \left\| W_{k,j}^{D*} a_k^{D,j} \right\|^2$, where $W_{k,j}^{D*} = S_{k,j} A_{k,j}^{\dagger}$, $S_{k,j}$ is a diagonal matrix of

base station j transmitted signal amplitudes for conventional channel k, $A_{k,j}$ is a column-wise concatenated matrix of known spatial signatures at base station j for conventional channel k, $A_{k,j}{}^\dagger$ represents the pseudoinverse of $A_{k,j}$, and $a_k^{D,j}$ is a downlink spatial signature for conventional channel k from base station j.

51. The method of claim 49 wherein the predicted second interference signal level for channel k is computed as $\sum_j \left\| \tilde{W}_{k,j}^{D\bullet} a_k^{D,j} \right\|^2$, where $\begin{bmatrix} \tilde{W}_{k,j}^{D\bullet} \\ w^\bullet \end{bmatrix} = \tilde{S}_{k,j}\left[A_{k,j} a_k^{D,j}\right]^\dagger$, $[\cdot]^\dagger$ represents a pseudoinverse operation, $\tilde{S}_{k,j}$ is a diagonal matrix of transmit signal amplitudes, $A_{k,j}$ is a column-wise concatenated matrix of known downlink spatial signatures at base station j for conventional channel k, $a_k^{D,j}$ is a downlink spatial signature for conventional channel k from base station j, and $w^\bullet$ is the bottom row of matrix $\tilde{S}_{k,j}\left[A_{k,j} a_k^{D,j}\right]^\dagger$.

52. The method of claim 42 further comprising a step for assigning a selected downlink conventional channel only if it is expected to result in an acceptable downlink received signal quality for each active channel based upon at least one downlink quality factor selected from a group consisting of: cost for new call; total base station transmitter power; predicted intermodulation distortion level; predicted interference-plus-noise level on each active channel; and predicted signal to interference-plus-noise ratio on each active channel.

53. The method of claim 52 wherein the predicted intermodulation distortion level is obtained by computing a crest factor for each downlink conventional channel, the crest factor value being indicative of an intermodulation distortion level that would result if a given downlink conventional channel were to be assigned.

54. A method for assignment of a full-duplex channel in which a duplex channel is selected from a set of duplex channels wherein each duplex channel of the set provides acceptable quality uplink communications, and the downlink channel assignment is made in accordance with existing rules of the system, the method comprising:

(a) creating a cost function that is indicative of a lack of communications quality expected from existing connections on each conventional uplink channel;

(b) computing a cost for each conventional uplink channel by using the cost function;

(c) assigning an uplink conventional channel by selecting a conventional uplink channel with a cost less than a prescribed cost level; and

(d) selecting a downlink channel in accordance with existing rules of the communication system.

55. The method of claim 54 wherein the full duplex channel assignment is made in accordance with the Personal Handyphone System (PHS) standard.

56. A method for assignment of a full-duplex channel in which a duplex channel is selected from a set of duplex channels wherein each duplex channel of the set provides acceptable quality downlink communications, and the uplink channel assignment is made in accordance with existing rules of the system, the method comprising:

(a) creating a cost function that is indicative of a lack of communications quality expected from existing connections on each conventional downlink channel;

(b) computing a cost for each conventional downlink channel by using the cost function;

(c) assigning an uplink conventional channel by selecting a conventional downlink channel with a cost less than a prescribed cost level; and

(d) selecting a uplink channel in accordance with existing rules of the communication system.

57. A method for assignment of a full-duplex channel in which a duplex channel is selected from a set of duplex channels wherein each duplex channel of the set provides acceptable quality downlink and uplink communications, the method comprising:

(a) creating a cost function that is indicative of a lack of communications quality expected from existing connections on each conventional duplex channel;

(b) computing a cost for each conventional duplex channel by using the cost function; and

(c) assigning a conventional duplex channel by selecting a conventional duplex channel with a cost less than a prescribed cost level.

58. A call admission control method, based on an interference cost function, for use in a wireless communication system for controlling admission of a connection for a new subscriber station, the method comprising:

(a) creating a cost function for evaluating the cost of interference-plus-noise on a conventional channel that has been selected for assignment;

(b) evaluating the cost function for obtaining a cost for a selected conventional channel if the selected conventional channel were to be assigned;

(c) comparing the cost with a prescribed channel assignment cost level; and

(d) admitting the call if the cost is less than the prescribed channel assignment cost level.

59. A call admission control method, based on a predictive interference model, for use in a wireless communication system for controlling admission of a connection for a new subscriber station, the method comprising:

(a) creating a model of the wireless communication system for predicting a received signal level and interference-plus-noise level on each conventional channel based upon existing connections;

(b) predicting the received signal level and interference-plus-noise level on each conventional channel based upon the model and upon existing connections; and

(c) selecting a conventional channel for the new subscriber station that has an acceptably high predicted signal-to-interference-plus-noise ratio (SINR); and

(d) admitting the call if the SINR is greater than a prescribed admission level SINR threshold value.

60. A call admission control method for use in a wireless communication system for controlling admission of a connection of a new subscriber station, the method, based on an estimate of existing system call load, comprising:

(a) estimating the existing call load for indicating how much of the system capacity is being utilized;

(b) prescribing a call load threshold that is indicative of a maximal call load allowed for the system; and

(c) comparing the existing call load with the call load threshold; and

(d) admitting a new call if the estimated existing call load is less than the call load threshold.

61. The method of claim 60 wherein estimating in step (a) further comprises monitoring the rate at which intracell handoffs occur, and estimating the existing call load from the rate at which intracell handoffs occur.

62. The method of claim 60 wherein step (a) further comprises estimating the existing call load by monitoring the rate of channel reassignments.

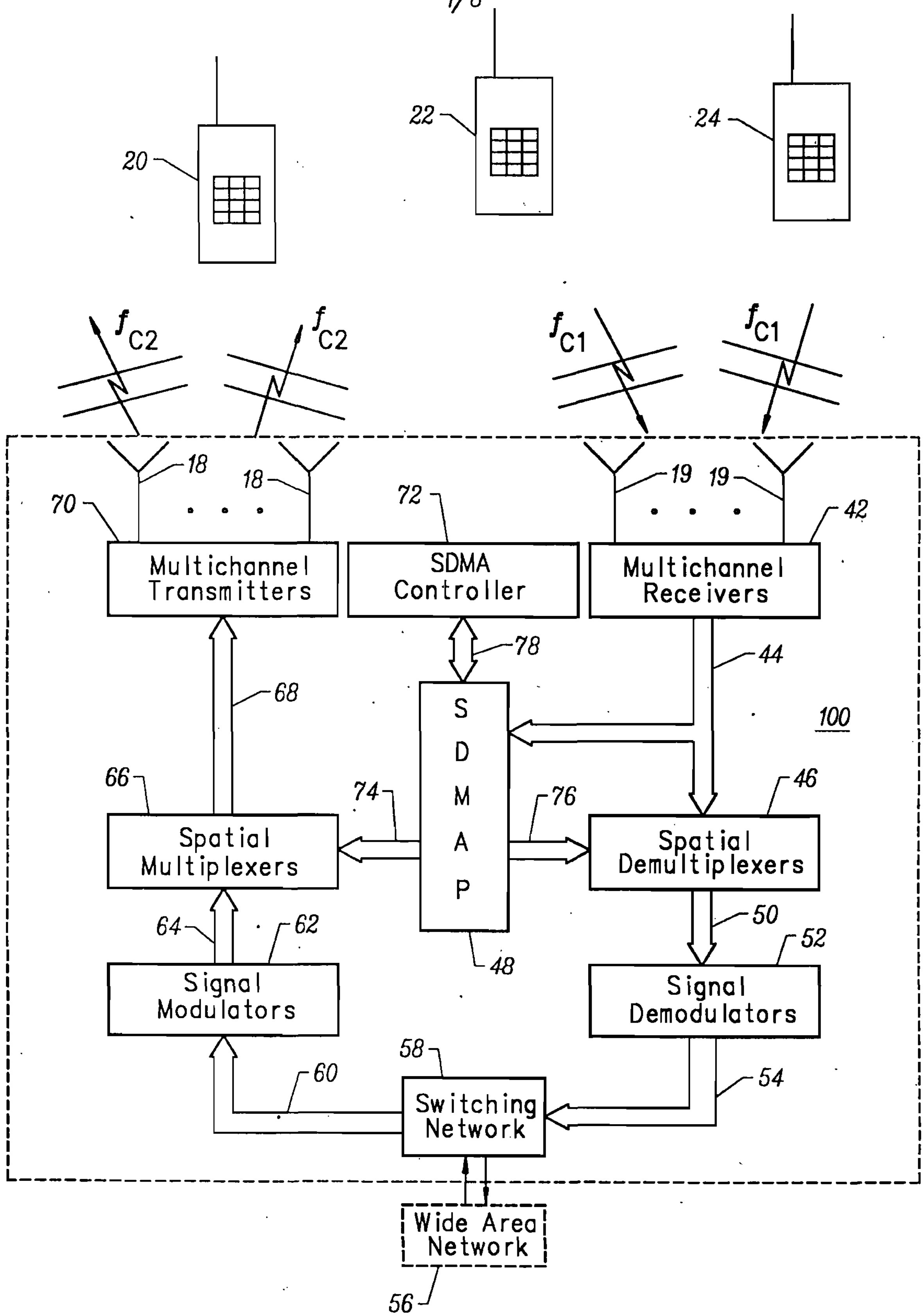
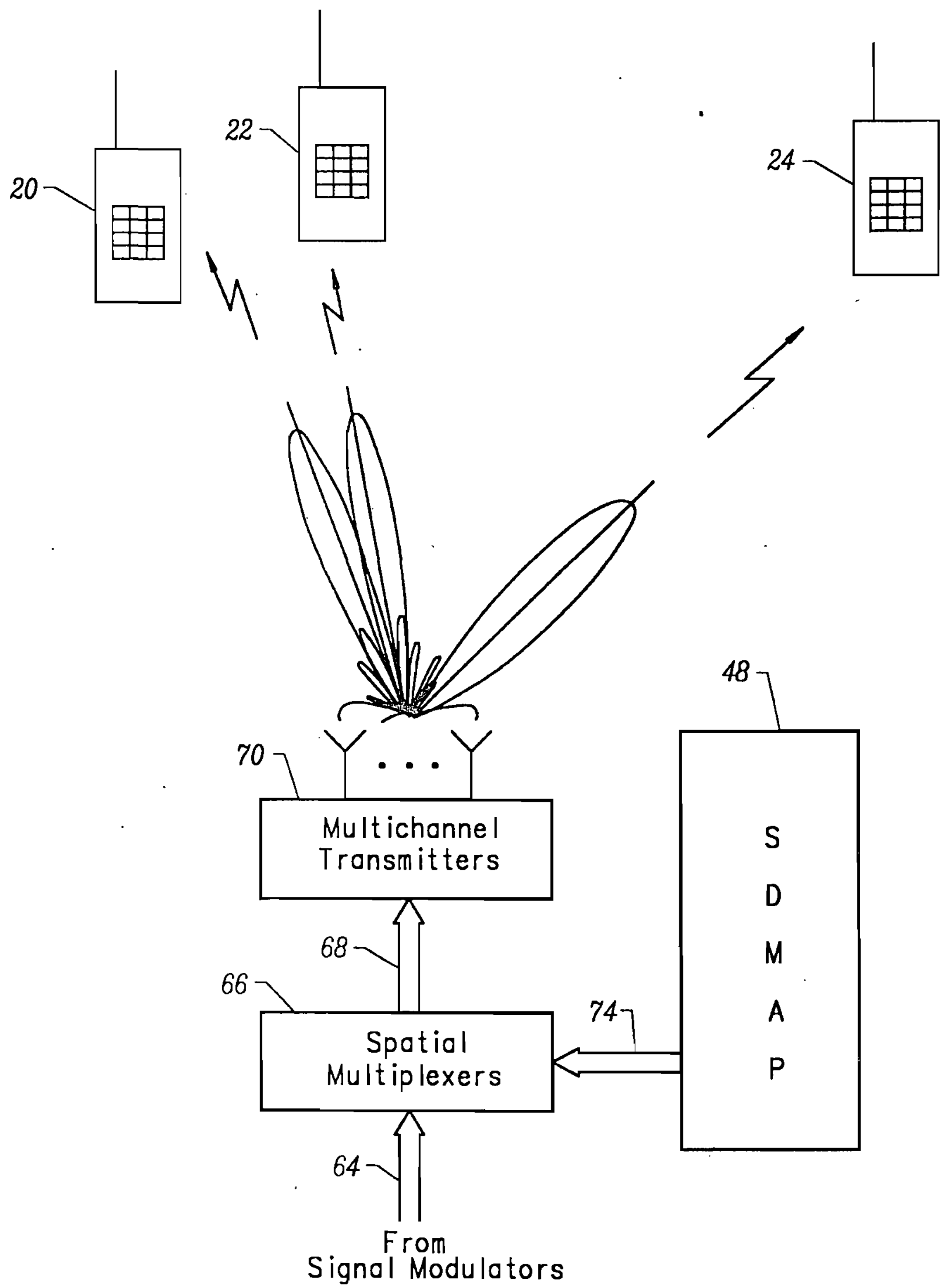

20
22
24
f_{C2}
f_{C2}
f_{C1}
f_{C1}
18
18
19
19
70
72
42
Multichannel Transmitters
SDMA Controller
Multichannel Receivers
78
44
68
100
S D M A P
66
74
76
46
Spatial Multiplexers
Spatial Demultiplexers
64
62
50
52
48
Signal Modulators
Signal Demodulators
58
60
54
Switching Network
Wide Area Network
56

FIG. 1



20
22
24
48
70
Multichannel Transmitters
68
66
Spatial Multiplexers
74
S D M A P
64
From Signal Modulators

FIG. 2

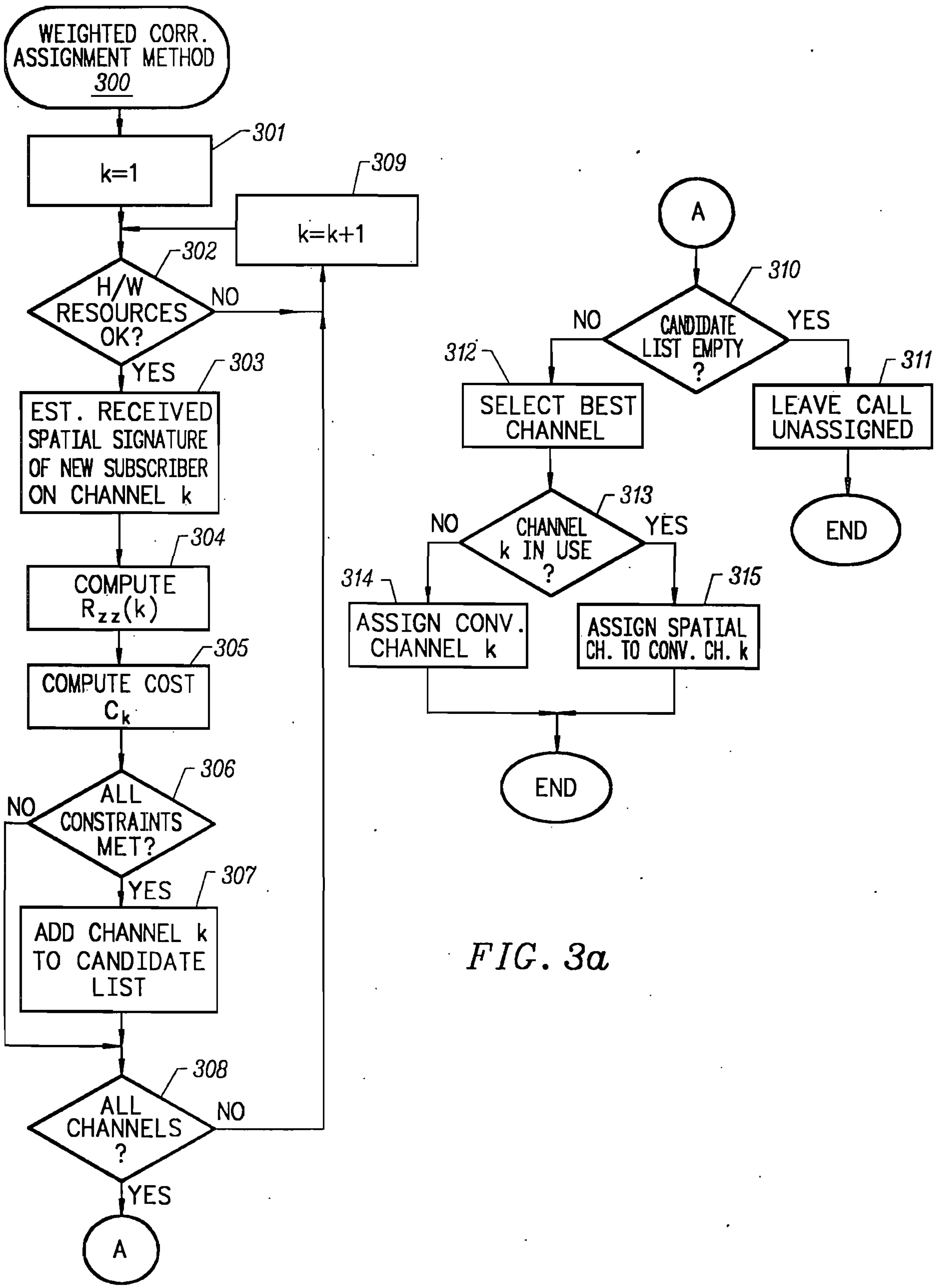

WEIGHTED CORR. ASSIGNMENT METHOD 300
301
k=1
309
k=k+1
302
H/W RESOURCES OK?
NO
YES
303
EST. RECEIVED SPATIAL SIGNATURE OF NEW SUBSCRIBER ON CHANNEL k
304
COMPUTE $R_{zz}(k)$
305
COMPUTE COST C_k
306
ALL CONSTRAINTS MET?
NO
YES
307
ADD CHANNEL k TO CANDIDATE LIST
308
ALL CHANNELS ?
NO
YES
A
A
310
CANDIDATE LIST EMPTY ?
NO
YES
312
SELECT BEST CHANNEL
311
LEAVE CALL UNASSIGNED
END
313
CHANNEL k IN USE ?
NO
YES
314
ASSIGN CONV. CHANNEL k
315
ASSIGN SPATIAL CH. TO CONV. CH. k
END

FIG. 3a

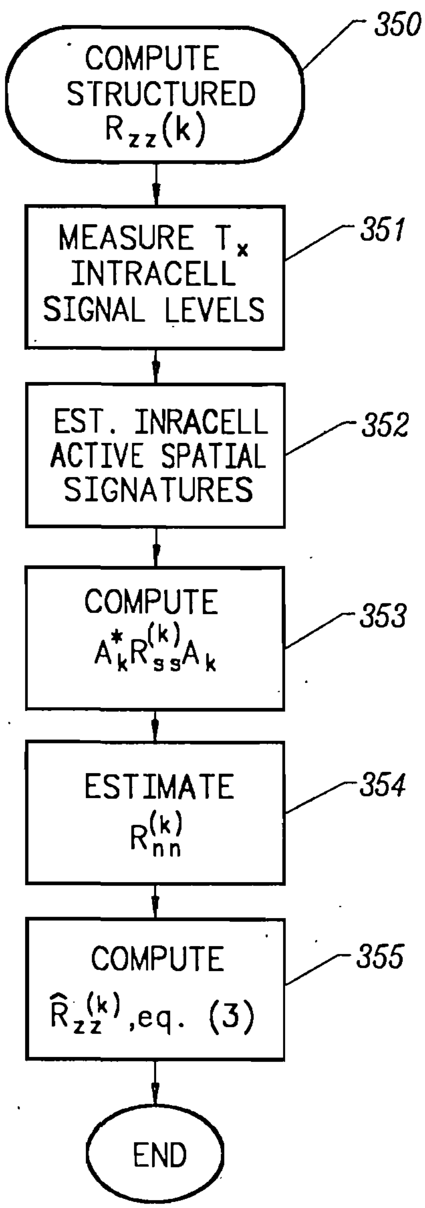

COMPUTE STRUCTURED $R_{zz}(k)$
350
MEASURE T_x INTRACELL SIGNAL LEVELS
351
EST. INRACELL ACTIVE SPATIAL SIGNATURES
352
COMPUTE $A_k^* R_{ss}^{(k)} A_k$
353
ESTIMATE $R_{nn}^{(k)}$
354
COMPUTE $\hat{R}_{zz}^{(k)}$,eq. (3)
355
END

FIG. 3c

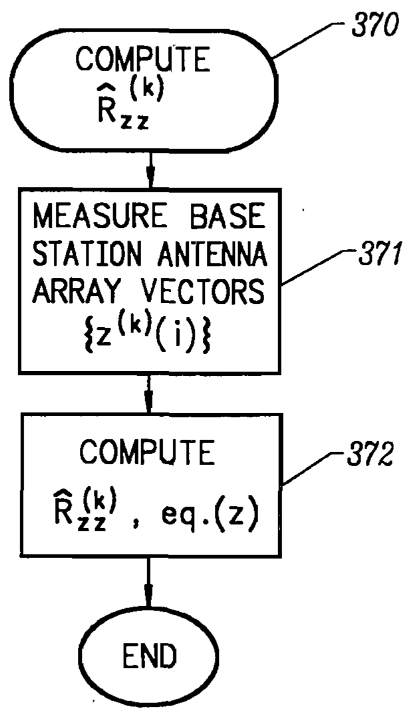

COMPUTE $\hat{R}_{zz}^{(k)}$
370
MEASURE BASE STATION ANTENNA ARRAY VECTORS $\{z^{(k)}(i)\}$
371
COMPUTE $\hat{R}_{zz}^{(k)}$, eq.(z)
372
END

FIG. 3b

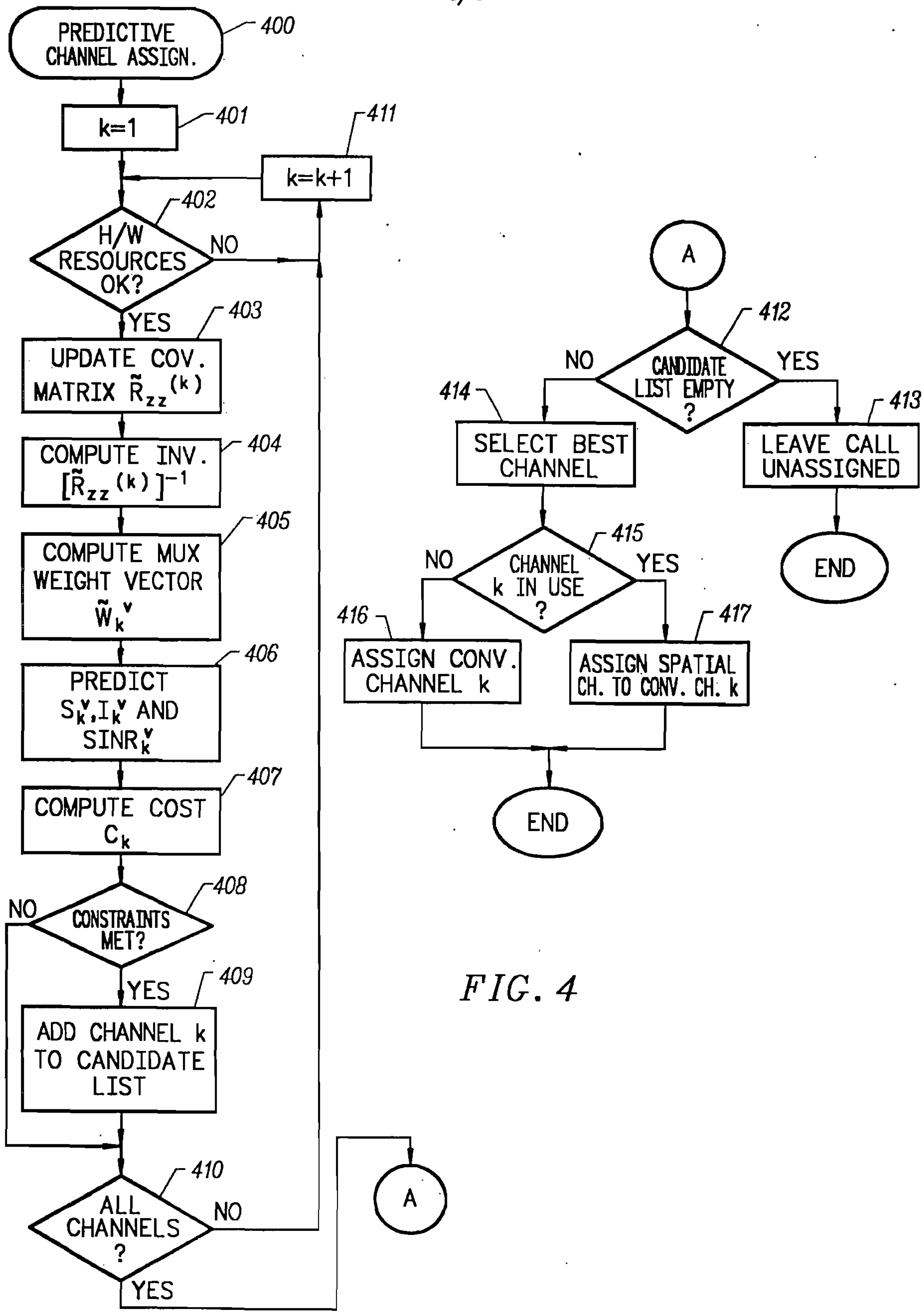

PREDICTIVE CHANNEL ASSIGN.
400
k=1
401
k=k+1
411
H/W RESOURCES OK?
402
NO
YES
UPDATE COV. MATRIX $\tilde{R}_{zz}^{(k)}$
403
COMPUTE INV. $[\tilde{R}_{zz}^{(k)}]^{-1}$
404
COMPUTE MUX WEIGHT VECTOR $\tilde{W}_k^v$
405
PREDICT S_k^v, I_k^v AND $SINR_k^v$
406
COMPUTE COST C_k
407
CONSTRAINTS MET?
408
NO
YES
ADD CHANNEL k TO CANDIDATE LIST
409
ALL CHANNELS ?
410
NO
YES
A
A
CANDIDATE LIST EMPTY ?
412
NO
YES
SELECT BEST CHANNEL
414
LEAVE CALL UNASSIGNED
413
END
CHANNEL k IN USE ?
415
NO
YES
ASSIGN CONV. CHANNEL k
416
ASSIGN SPATIAL CH. TO CONV. CH. k
417
END

FIG. 4

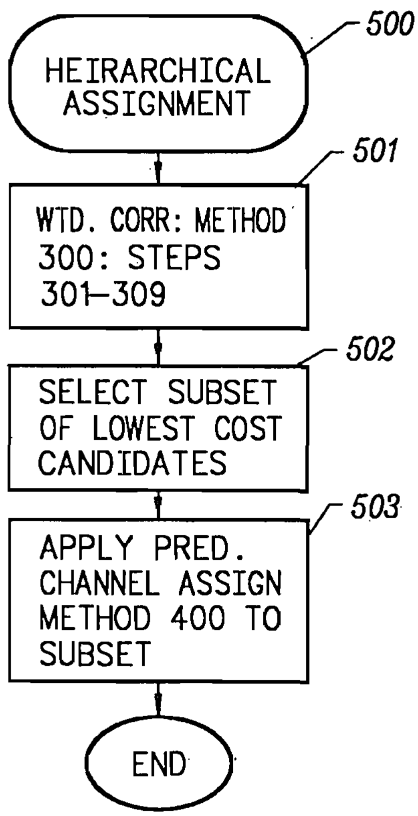

500
HEIRARCHICAL ASSIGNMENT
501
WTD. CORR: METHOD 300: STEPS 301-309
502
SELECT SUBSET OF LOWEST COST CANDIDATES
503
APPLY PRED. CHANNEL ASSIGN METHOD 400 TO SUBSET
END

FIG. 5

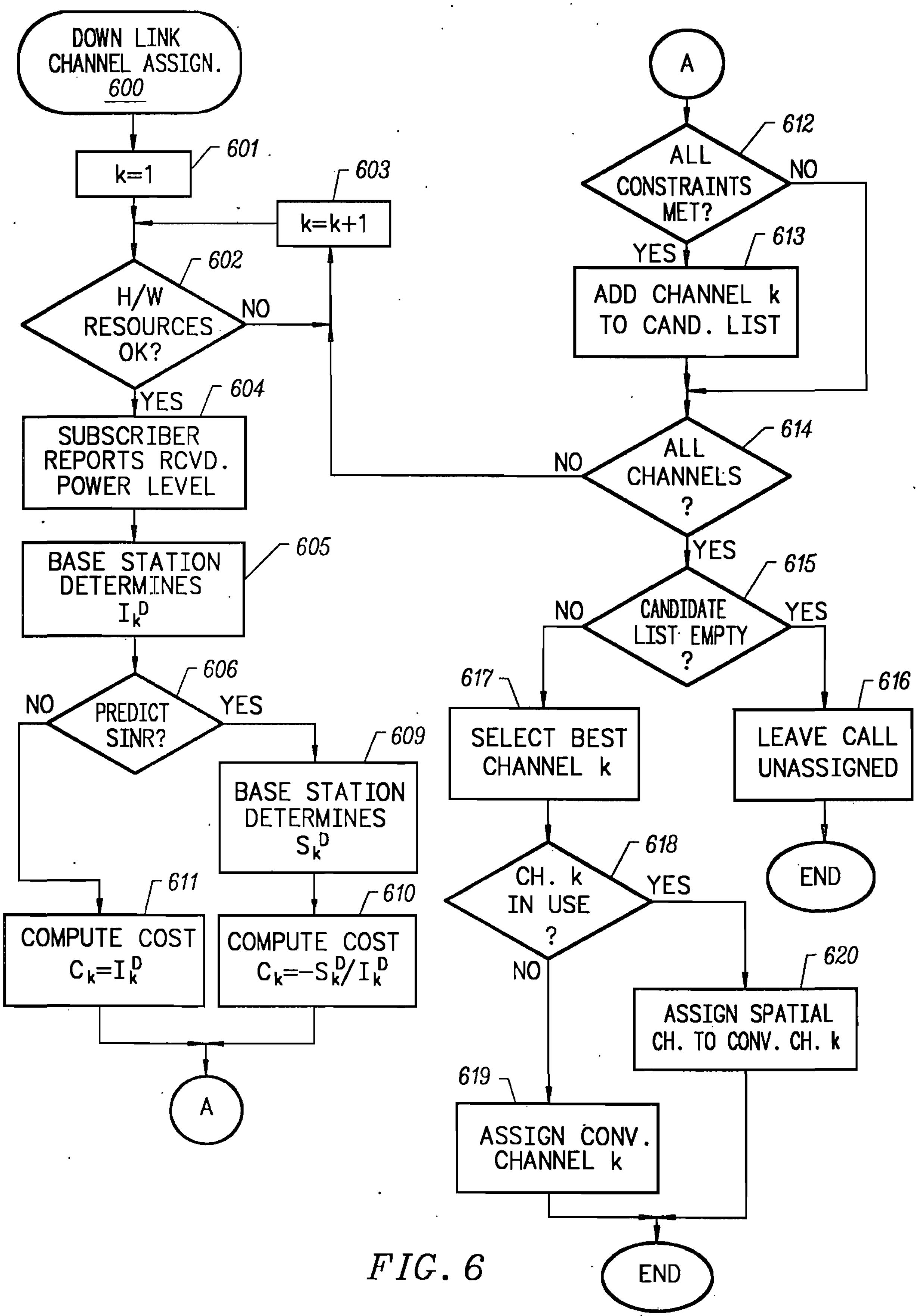
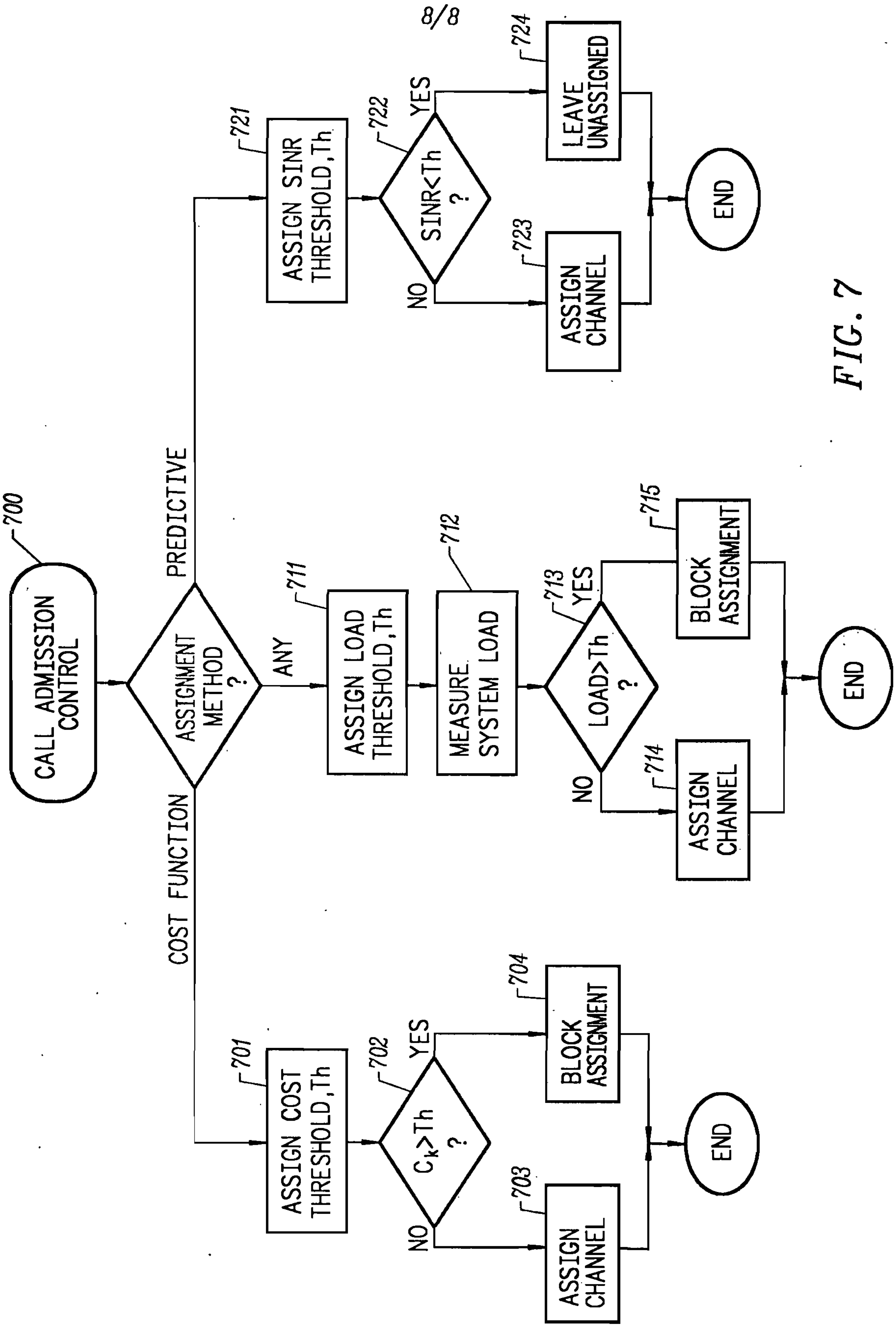

DOWN LINK CHANNEL ASSIGN. 600
k=1
601
603
k=k+1
602
H/W RESOURCES OK?
NO
YES
604
SUBSCRIBER REPORTS RCVD. POWER LEVEL
605
BASE STATION DETERMINES I_k^D
606
PREDICT SINR?
NO
YES
609
BASE STATION DETERMINES S_k^D
611
COMPUTE COST $C_k=I_k^D$
610
COMPUTE COST $C_k=-S_k^D/I_k^D$
A
A
612
ALL CONSTRAINTS MET?
NO
YES
613
ADD CHANNEL k TO CAND. LIST
614
ALL CHANNELS ?
NO
YES
615
CANDIDATE LIST EMPTY ?
NO
YES
617
SELECT BEST CHANNEL k
616
LEAVE CALL UNASSIGNED
END
618
CH. k IN USE ?
YES
NO
620
ASSIGN SPATIAL CH. TO CONV. CH. k
619
ASSIGN CONV. CHANNEL k
END

FIG. 6



700
CALL ADMISSION CONTROL
ASSIGNMENT METHOD ?
PREDICTIVE
ANY
COST FUNCTION
721
ASSIGN SINR THRESHOLD,Th
722
SINR<Th ?
YES
NO
723
ASSIGN CHANNEL
724
LEAVE UNASSIGNED
END
711
ASSIGN LOAD THRESHOLD,Th
712
MEASURE SYSTEM LOAD
713
LOAD>Th ?
YES
NO
714
ASSIGN CHANNEL
715
BLOCK ASSIGNMENT
END
701
ASSIGN COST THRESHOLD,Th
702
C_k>Th ?
YES
NO
703
ASSIGN CHANNEL
704
BLOCK ASSIGNMENT
END

FIG. 7

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US97/23731

A. CLASSIFICATION OF SUBJECT MATTER
IPC(6) :H04Q 7/00
US CL : 370/329; 455/63
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED
Minimum documentation searched (classification system followed by classification symbols)
U.S. : Please See Extra Sheet.

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X	US 5,093,924 A (TOSHIYUKI ET AL) 03 March 1992, col. 6, lines 8-53.	1-4, 11-18, 22-23, 30-35, 42-43, 45-47, and 54-59
X	US 5,530,917 A (ANDERSSON ET AL) 25 June 1996, col. 2, line 38 to col. 3, line 54.	60

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

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Date of the actual completion of the international search
29 MARCH 1998

Date of mailing of the international search report
05 MAY 1998

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INTERNATIONAL SEARCH REPORT

International application No.
PCT/US97/23731

B. FIELDS SEARCHED
Minimum documentation searched
Classification System: U.S.

370/329, 252, 276, 277, 310, 328, 329, 331, 332, 333, 334, 341, 431, 465, 468; 455/63, 423, 424, 425, 434, 450, 451, 452, 453, 464, 509, 510, 67.1

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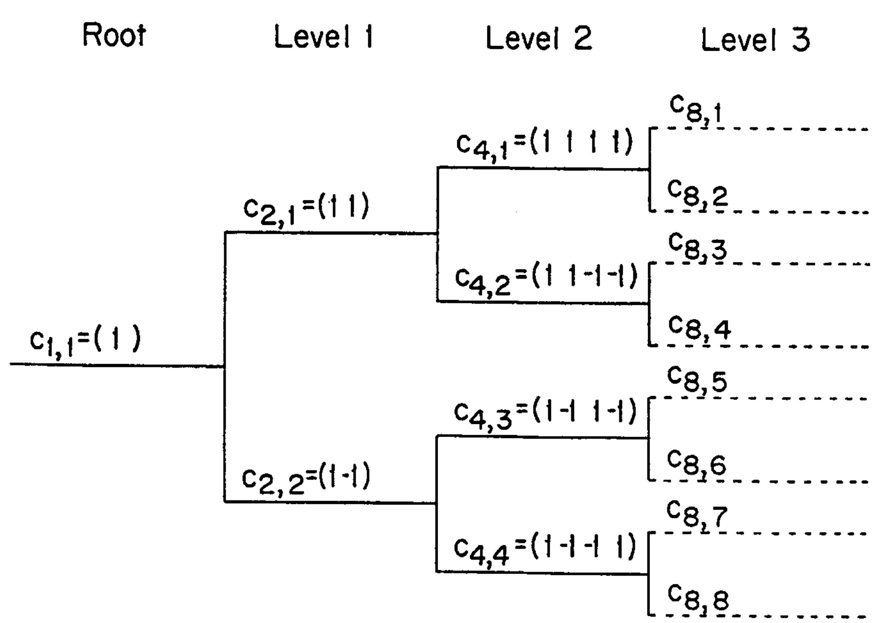
(74) Agent: ERICSSON RADIO SYSTEMS AB; Common Patent Dept., S-164 80 Stockholm (SE).

(81) Designated States: AL, AM, AT, AU, AZ, BA, BB, BG, BR, BY, CA, CH, CN, CU, CZ, DE, DK, EE, ES, FI, GB, GE, GH, GM, GW, HR, HU, ID, IL, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MD, MG, MK, MN, MW, MX, NO, NZ, PL, PT, RO, RU, SD, SE, SG, SI, SK, SL, TJ, TM, TR, TT, UA, UG, UZ, VN, YU, ZW, ARIPO patent (GH, GM, KE, LS, MW, SD, SZ, UG, ZW), Eurasian patent (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European patent (AT, BE, CH, CY, DE, DK, ES, FI, FR, GB, GR, IE, IT, LU, MC, NL, PT, SE), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, ML, MR, NE, SN, TD, TG).

Published
With international search report.

(54) Title: CHANNELIZATION CODE ALLOCATION FOR RADIO COMMUNICATION SYSTEMS



(57) Abstract

Variable spreading factors and multi-code transmissions are flexibly accomodated by assigning spreading codes in accordance with the described techniques. Spreading codes are assigned so that the control channel is orthogonal to all physical channels in the composite spread spectrum signal. Power balance between in-phase (I)and quadrature (Q) branches in the transmitter is also provided by assigning physical channels to appropriate branches and splitting physical channels, where necessary.

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CHANNELIZATION CODE ALLOCATION FOR RADIO COMMUNICATION SYSTEMS

BACKGROUND

This invention generally relates to variable data rate transmissions and, more particularly, to techniques for efficiently allocating spreading codes for variable rate data transmissions.

Cellular radio communication systems have recently been developed that use spread spectrum modulation and code division multiple access (CDMA) techniques. In a typical direct sequence CDMA system, an information data stream to be transmitted is superimposed on a much-higher-symbol-rate data stream sometimes known as a spreading sequence. Each symbol of the spreading sequence is commonly referred to as a chip. Each information signal is allocated a unique spreading code that is used to generate the spreading sequence typically by periodic repetition. The information signal and the spreading sequence are typically combined by multiplication in a process sometimes called coding or spreading the information signal. A plurality of spread information signals are transmitted as modulations of radio frequency carrier waves and are jointly received as a composite signal at a receiver. Each of the spread signals overlaps all of the other coded signals, as well as noise-related signals, in both frequency and time. By correlating the composite signal with one of the unique spreading sequences, the corresponding information signal can be isolated and decoded.

As radiocommunication becomes more widely accepted, it will be desirable to provide various types of radiocommunication services to meet consumer demand. For example, support for facsimile, e-mail, video, internet access, etc. via radiocommunication systems is envisioned. Moreover, it is expected that users may wish to access different types of services at the same time. For example, a video conference between two users would involve both speech and video support. Some of these different services will require relatively high data rates compared with speech service that has been conventionally supplied by radio communication systems, while other services will require variable data rate service. Thus, it is anticipated that future

radio communication systems will need to be able to support high data rate communications as well as variable data rate communications.

Several techniques have been developed to implement variable rate communications in CDMA radio communication systems. From the perspective of transmitting data at varying rates, these techniques include, for example, discontinuous transmission (DTX), variable spreading factors, multi-code transmission and variable forward error correction (FEC) coding. For systems employing DTX, transmission occurs only during a variable portion of each frame, i.e., a time period defined for transmitting a certain size block of data. The ratio between the portion of the frame used for transmission and the total frame time is commonly referred to as the duty cycle γ. For example, when transmitting at the highest possible rate, i.e., during the entire frame period, $\gamma = 1$, while for zero rate transmissions, e.g., during a pause in speech, $\gamma = 0$. DTX is used, for example, to provide variable data rate transmissions in systems designed in accordance with the U.S. standard entitled "Mobile Station-Base Station Compatibility Standard for Dual-Mode Wideband Spread Spectrum Cellular System", TIA/EIA Interim Standard TIA/EIA/IS-95 (July 1993) and its revision TIA/EIA Interim Standard TIA/EIA/IS-95-A (May 1995). Such standards that determine the features of U.S. cellular communication systems are promulgated by the Telecommunications Industry Association and the Electronic Industries Association located in Arlington, Virginia.

Varying the spreading factor is another known technique for providing variable data rate communication. As mentioned above, DS-CDMA spread spectrum systems spread data signals across the available bandwidth by multiplying each of the data signals with spreading sequences. By varying the number of chips per data symbol, i.e., the spreading factor, while keeping the chip rate fixed, the effective data rate can be controllably varied. In typical implementations of the variable spreading factor approach, the spreading factor is limited by the relationship to $SF = 2^k x\ SF_{min}$ where SF_{min} is the minimum allowed spreading factor corresponding to the highest allowed user rate.

Another known technique for varying the transmitted data rate is commonly referred to as multi-code transmission. According to this technique, data is transmitted using a variable number of spreading codes where the exact number of codes used depends on the instantaneous user bit rate. Effectively, this means allocating a variable number of physical channels to a connection to provide variable bandwidth. An example of multi-code transmission is described in U.S. Patent Application Serial No. 08/636,648 entitled "Multi-Code Compressed Mode DS-CDMA Systems and Methods", filed on April 23, 1996, the disclosure of which is incorporated here by reference.

Yet another technique for varying the transmitted data rate in radio communication systems involves varying the FEC. More specifically, the rate of the forward error correction (FEC) coding is varied by using code-puncturing and repetition or by switching between codes of different rates. In this way the user rate is varied while the channel bit rate is kept constant. Those skilled in the art will appreciate the similarities between varying the FEC and a variable spreading factor as mechanisms to implement variable rate transmission.

In both the uplink and downlink, it is desirable that any number of logical channels can be transmitted simultaneously to support a single connection between a base station and a mobile station to support various data rates. To transmit these logical channels over the radio interface, a number of physical channels are allocated. These physical channels are separated using different spreading codes (channelization codes), i.e., multicode transmission is used. Each physical channel can have one of several possible data rates, i.e., one of several possible spreading factors is used when spreading the data transmitted on the physical channel. To date, however, a flexible solution which allocates code words to physical channels taking into consideration the codes which have already been allocated to other channels and power considerations associated with the in-phase (I) and quadrature (Q) transmitter branches has not been provided.

Accordingly, it would be desirable to create new techniques and systems for allocating spreading codes in a flexible manner that supports multicode transmissions and variable spreading factors, and that optimizes power efficiency.

SUMMARY

These and other problems associated with previous communication systems are solved by Applicants' invention, wherein spreading codes are allocated for physical channels taking into consideration the spreading codes already allocated to other physical channels to be transmitted in parallel therewith. For example, if the physical channel being allocated a spreading code is a control channel (PCCH), then techniques according to the present invention investigate whether another physical channel on either the I or Q branches of the transmitter has already been assigned a spreading code so that the PCCH can be allocated a spreading code which makes the PCCH orthogonal to all other physical channels used in the composite spread spectrum signal. Moreover, for physical data channels (PDCH), techniques according to the present invention determine if any other channels have previously been assigned spreading codes on the same I or Q branch as the channel currently under investigation. If so, this PDCH is allocated a spreading code that makes the PDCH orthogonal to other PDCHs in the same branch, as well as to the PCCH.

According to other exemplary embodiments of the present invention, in addition to assigning a spreading code to each physical channel, the physical channels are also allocated between the I and Q branches of the transmitter in a manner intended to balance power between the two branches and improve power amplifier performance. For example, if the data rate associated with a connection to be set up is relatively low, then the connection may be supported by one PDCH and one PCCH, one of which is assigned to the I branch of the transmitter and the other to the Q branch. If, however, the data rate associated with a connection to be set up is relatively high, then assigning the PDCH to one branch and the PCCH to the other creates a large power discrepancy between the two branches. In such a case, the data can be transmitted on two PDCHs

each of which are allocated to the I and Q branches of the transmitter, respectively, and the control channel can be allocated to either the I or Q branch.

BRIEF DESCRIPTION OF THE DRAWINGS

The features and objects of Applicants' invention will be understood by reading this description in conjunction with the drawings, in which:

FIG. 1A is a block diagram representation of an exemplary transmitter structure in which the present invention can be implemented;

FIG. 1B illustrates an alternative scrambling technique which can be implemented in the transmitter of FIG. 1A;

FIG. 2 is an exemplary code tree;

FIG. 3 is a flowchart depicting allocation of physical channels between the I and Q branches of a transmitter according to an exemplary embodiment of the present invention; and

FIG. 4 is a flowchart illustrating the allocation of spreading codes to physical channels according to the present invention.

DETAILED DESCRIPTION

While this description is written in the context of cellular communications systems involving portable or mobile radio telephones, it will be understood by those skilled in the art that Applicants' invention may be applied to other communications applications.

According to exemplary embodiments of the present invention, CDMA systems can support variable bit rate services, such as speech, by providing control information in each frame which specifies the instantaneous data symbol rate for that frame. In order to accomplish this in a regular time interval, physical channels can be organized in frames of equal length (timewise). Each frame carries an integer number of chips and an integer number of information bits.

Using this exemplary frame structure, bit rate control information can be provided for every CDMA frame by transmitting this information on a separate

physical channel. The physical channels carrying the data and the control information (e.g., including pilot/reference symbols for channel estimation, power control commands and rate information of the data) can be denoted as physical data channel (PDCH) and physical control channel (PCCH), respectively. Each connection between a mobile station and a base station will be supported by a PCCH and at least one PDCH. The spreading code, symbol rate, or equivalently spreading factor, of the PCCH are known a priori to the receiver. In this way, the receiver can determine the data rate of the PDCH(s) from the PCCH prior to demodulating/decoding the PDCH(s). Exemplary techniques for handling BRI information are described in commonly-assigned, copending U.S. Patent Application Serial No. ________________, entitled "Low-Delay Rate Detection for Variable Rate Communication Systems" to Dahlman et al., filed on an even date herewith.

Many potential advantages are attributable to variable rate transmission. For example, interference can be reduced for various users of the system since the chip rate is kept constant and a lower bit rate gives a higher spreading factor, thus allowing a lower transmit power. Those skilled in the art will readily appreciate how this ability to vary the information rate in a CDMA system can be used advantageously to vary other parameters. However, techniques for efficiently allocating spreading codes to the various physical channels (i.e., PCCH and PDCH(s)) are needed and described below.

A physical channel is a bit stream of a certain rate, that is spread using a certain code and allocated to either the in-phase (I) or quadrature (Q) branch in a transmitter. Variable rate services are supported through spreading with a variable spreading factor as described above. A number of data streams are spread to the chip rate using Walsh codes of different length, followed by summation and, if desired, scrambling. The structure of an exemplary transmitter (usable, e.g., in either a base station or a mobile station) which performs these spreading, summing and scrambling operations is illustrated in Figure 1A.

Therein, a first data stream I_1 is supplied to multiplier 10 having a data rate of R_I which is equal to the chip rate R_c divided by the spreading factor SF_{I1} for that data stream. This data stream is spread with a channelization code word C_{I1} having a length

of 2^k chips which is selected such that the output of multiplier 30 has a chip rate R_c by selecting a value for k that is related to the desired data rate of physical channel I_1. For example, a physical channel data rate of 250 kbps is spread to a chip rate of 4 Mcps by using a channelization code of 16 (2^4) chips long. More details regarding the allocation of a particular channelization code according to the present invention are described below. Similarly, additional data streams are supplied to multipliers 12, 14 and 16 (and other branches which are unillustrated) to spread their respective data streams with channelization code words having a length which is selected to result in a chip rate R_c. The rate of the data streams can be limited to such an interval that the spreading factors used are larger or equal to a predetermined SF_{min}. Each physical channel is then weighted by respective amplifiers 18, 20, 22 and 24. The weights can be individually chosen to allocate power to each physical channel so that predetermined quality requirements, e.g., the bit error rate of each physical channel, are satisfied. The physical channels in the "I" branch of the transmitter are summed at summer 26. Similarly, the physical channels in the "Q" branch of the transmitter are summed at summer 28. Scrambling, if desired, is then performed on the superimposed physical channels. This can be done in at least two ways. First, as shown in Figure 1A, scrambling can be performed by forming the I and Q pairs as a complex number at blocks 30 and 32 and then multiplying the result with another complex number (i.e., the complex valued scrambling code $c_{scramb}=c_I+jc_Q$) at block 34. Scrambling can also be performed on the I and Q branches separately as illustrated in Figure 1B, by multiplying I and Q with two real valued scrambling codes c_I and c_Q at blocks 36 and 38. The scrambling code is clocked at the chip rate. The resultant signal is output, e.g., to transmit signal processing circuitry (e.g. a QPSK or O-QPSK modulator followed by, possibly, pulse-shaping filters), amplified by a transmit power amplifier (not shown) and ultimately coupled to an antenna (also not shown).

The Walsh codes used for spreading at multipliers 10-16 can be viewed in a tree like manner, as illustrated in Figure 2. Codes on the same level in the tree are orthogonal and have the same spreading factor. Thus, codes $c_{4,1}$, $c_{4,2}$, $c_{4,3}$ and $c_{4,4}$ are orthogonal codes each of which have the same spreading factor, i.e., the same number

of chips. If a physical channel is spread with a first code in the tree, and another physical channel is spread with another code which is (1) not the same as the first code, (2) not to the left of the first code on the path to the root of the tree and (3) not in the subtree which has the first code as the root, then the two spread physical channels will be orthogonal. For example, if the PCCH is allocated code $c_{4,1}$ and a PDCH is allocated code $c_{8,5}$, then these two spread channels would be orthogonal. If, however, the PDCH was allocated code $c_{8,1}$ or $c_{8,2}$, then the PCCH and PDCH would be non-orthogonal. Every physical channel is allocated a spreading code from the tree, with spreading factors matching the respective data rates. As the data rate varies for a particular PDCH, a code from a different level of the tree will be allocated. For example, increasing data rates will cause code selection to move to the left in the tree, while for decreasing data rates code selection will move to the right. Thus, a typical variable rate PDCH will typically move up and down along a certain path in the code tree as its data rate varies. Allocation of physical channels to the I and Q branches of the transmitter, as well as codes from the code tree in Figure 2 as spreading codes (e.g., c_{I1}, c_{Q1}, etc. in Figure 1A) can be made according to the following rules in accordance with the present invention.

Figure 3 is a flowchart which illustrates an exemplary technique for allocating the physical channels between the I and Q branches of a transmitter according the present invention for the case where a single PDCH can be used (i.e., has sufficient bandwidth) to support a connection. Those skilled in the art will appreciate that this technique provides for a relatively balanced transmit power for the each of the I and Q branches which in turn provides better power amplifier performance. The flow begins at block 40 wherein it is determined whether the power that would be needed to transmit the single PDCH is significantly greater than that needed to transmit the PCCH. For example, if the PDCH is to be transmitted at a much higher rate than the PCCH or if the quality of service (QoS) requirements for the PDCH are higher, then the power requirements will be correspondingly higher. In such a case, the flow proceeds to block 42 wherein the data stream is split into two lower rate PDCHs. The three physical channels can then be allocated, for example, as illustrated in block 42 to

the I and Q branches in a manner which will help to more evenly balance the transmit power between these two branches. If, on the other hand, it is determined at block 40 that the PDCH is not to be transmitted at a significantly greater power than the PCCH, then the flow proceeds to block 44 wherein the control channel is allocated to one of the branches and the data channel to the other. Note that the particular selection of Q and I in blocks 42 and 44 is exemplary only and that these designations could of course be reversed.

Having assigned the physical channels to a respective one of the I and Q branches in the transmitter, the next allocation to be made according to the present invention is the selection of a spreading code for each of the physical channels. According to the present invention, the spreading code selected to spread the PCCH should be such that the PCCH is orthogonal to all of the other physical channels to be transmitted in the composite spread spectrum signal, i.e., orthogonal to all channels in both the I and Q branches. This characteristic is desirable because the PCCH must first be demodulated and decoded at the receiver to provide channel estimates which are used to process the data channels transmitted in the same spread spectrum signal. Accordingly, an exemplary technique for allocating spreading codes according to the present invention will now be described with respect to the flowchart of Figure 4. The flow begins at block 52 wherein it is determined whether the present channel that is being allocated a spreading code is a data channel or a control channel. If the channel currently being allocated a spreading code is a PDCH then the flow proceeds to block 54. Therein, this PDCH is allocated a spreading code which makes the PDCH orthogonal to the PCCH (if the PCCH has already been allocated a spreading code) and which makes the PDCH orthogonal to any other PDCH that is on the same I or Q branch of the transmitter. For example, suppose that at the time this particular PDCH is being allocated a spreading code that the PCCH has already been allocated code $c_{4,1}$ and another PDCH has already been allocated code $c_{8,5}$. Further, assume that this particular PDCH is to be transmitted at a data rate that requires a level 3 code with respect to the code tree of Figure 2. According to the present invention, this particular PDCH could then be allocated any of codes $c_{8,3}$, $c_{8,4}$, $c_{8,6}$, $c_{8,7}$ and $c_{8,8}$. This PDCH

could not be allocated to codes $c_{8,1}$ or $c_{8,2}$ since such allocations would result in non-orthogonality with the control channel. This PDCH could, however, be allocated code $c_{8,5}$ if it is assigned to the opposite transmitter branch of the PDCH which has already been assigned this spreading code.

The flow then proceeds to block 56 whereupon more codes are allocated if additional channels remain. Otherwise the process terminates. If, at block 52, a control channel is being evaluated for spreading code allocation, then the flow proceeds to block 58. Therein, a code is selected which makes the control channel orthogonal to all channels previously allocated codes so that the PCCH can be readily decoded and demodulated at the receiver to provide channel estimates for use and evaluating the data channels.

It will be understood that Applicants' invention is not limited to the particular embodiments described above and that modifications may be made by persons skilled in the art. The scope of Applicants' invention is determined by the following claims, and any and all modifications that fall within that scope are intended to be included therein.

We Claim:

1. A transmitter having an in-phase (I) branch and a quadrature (Q) branch for transmitting a composite, spread spectrum signal including at least two physical channels, said transmitter comprising:

means, associated with said I branch, for spreading data associated with one of said at least two physical channels using a first spreading code to generate a first spread physical channel; and

means, associated with said Q branch, for spreading data associated with another of said at least two physical channels using a second spreading code to generate a second spread physical channel;

wherein said first and second spreading codes have a different number of chips and said first and second spreading codes are selected so that said first and second spread physical channels are orthogonal to one another.

2. The transmitter of claim 1, wherein said one of said at least two physical channels is a control channel (PCCH) and said another of said at least two physical channels is a data channel (PDCH).

3. The transmitter of claim 2, further comprising:

means for balancing power associated with said I and Q branches of said transmitter by selectively allocating said at least two physical channels to said I and Q branches based on transmit power requirements.

4. The transmitter of claim 3, wherein said at least two physical channels include a second PDCH which is spread using a third code to generate a third spread physical channel, and wherein said means for balancing power allocates said second PDCH to a same branch of said transmitter as said PCCH to based on said transmit power requirement.

5. The transmitter of claim 4, wherein said second and third spread physical channels are orthogonal.

6. The transmitter of claim 4, wherein said second and third spread physical channels are non-orthogonal.

7. The transmitter of claim 4, wherein said second and third codes are the same codes.

8. A method for allocating spreading codes to a plurality of physical channels to be transmitted in a composite spread spectrum signal in a radio communication system comprising the steps of:

allocating a first spreading code having a first number of chips to a control channel so that said control channel is orthogonal to others of said plurality of physical channels in said composite spread spectrum signal; and

allocating a second spreading code having a second number of chips different from said first number of chips to a first data channel, which second spreading code is selected such that said control channel and said first data channel are orthogonal to one another.

9. The method of claim 8, wherein said control channel conveys reference information usable to make channel estimates.

10. The method of claim 8 further comprising the step of:

allocating a third spreading code having a third bit length to a second data channel, said third spreading code selected such that said control channel and said second data channel are orthogonal to one another.

11. The method of claim 10, wherein said first and second data channels are orthogonal.

12. The method of claim 10, wherein said first and second data channels are non-orthogonal.

13. The method of claim 10, wherein said second and third spreading codes are the same codes.

14. The method of claim 10, further comprising the steps of:

assigning said second data channel to one of an I and a Q branch in a transmitter; and

assigning said third data channel to the other of said I and Q branches.

15. The transmitter of claim 1, further comprising:

means for scrambling said first and second spread physical channels of said I and Q branches,

16. A method for allocating spreading codes to a plurality of physical channels to be transmitted in a composite spread spectrum signal in a radio communication system comprising the steps of:

allocating a first spreading code having a first number of chips to a first data channel; and

allocating a second spreading code having a second number of chips different from said first number of chips to a control data channel, which second spreading code is selected such that said control channel and said first data channel are orthogonal to one another.

17. The method of claim 16, wherein said control channel conveys reference information usable to make channel estimates.

18. The method of claim 16 further comprising the step of:

allocating a third spreading code having a third bit length to a second data channel, said third spreading code selected such that said control channel and said second data channel are orthogonal to one another.

19. The method of claim 18, wherein said first and second data channels are orthogonal.

20. The method of claim 18, wherein said first and second data channels are non-orthogonal.

21. The method of claim 18, wherein said second and third spreading codes are the same codes.

22. The method of claim 18, further comprising the steps of:

assigning said second data channel to one of an I and a Q branch in a transmitter; and

assigning said third data channel to the other of said I and Q branches.

Fig. 1A

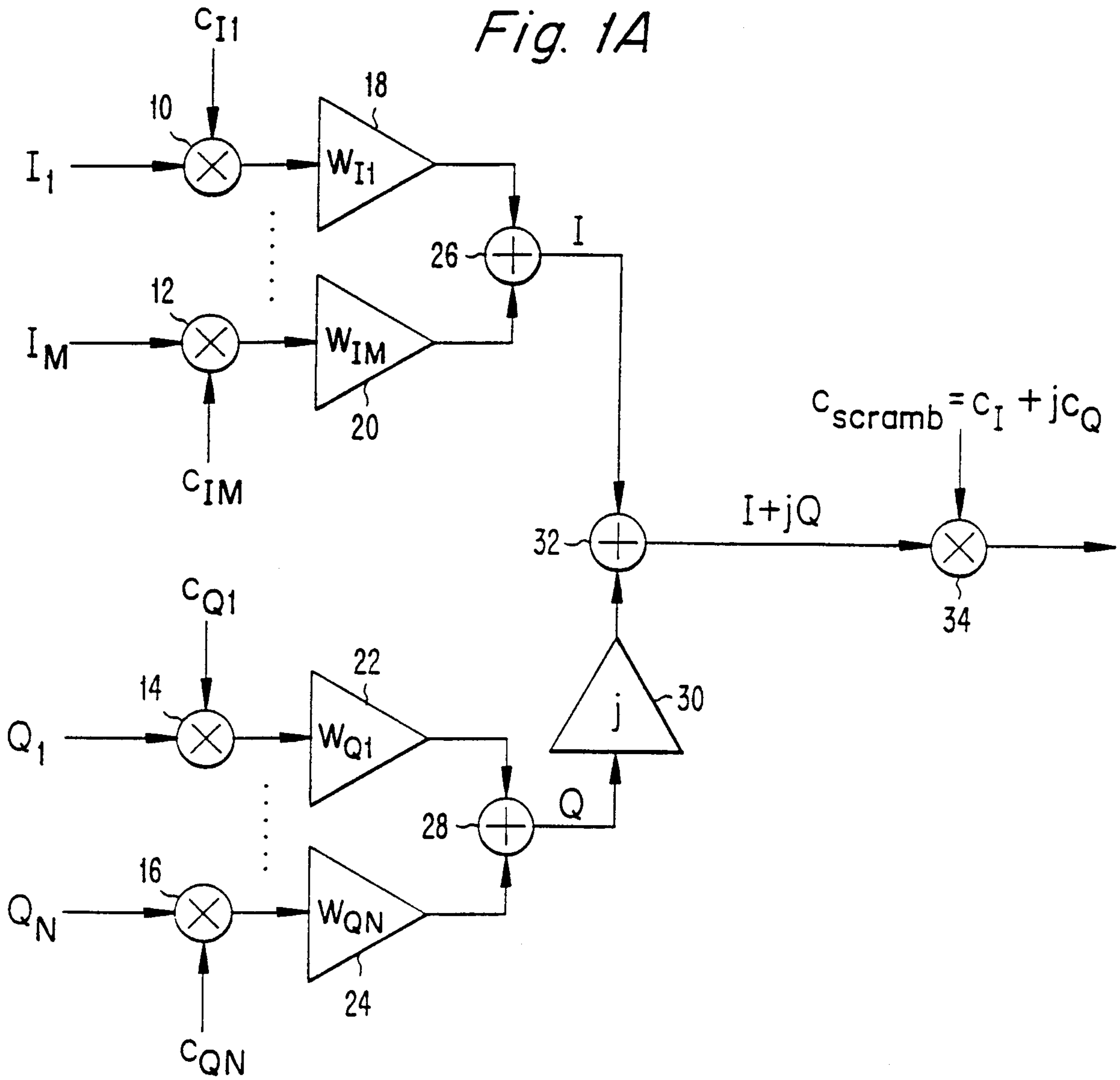

c_{I1}
10
18
I_1
W_{I1}
26
I
12
I_M
W_{IM}
20
c_{IM}
$c_{scramb} = c_I + jc_Q$
32
I+jQ
34
c_{Q1}
14
22
Q_1
W_{Q1}
j
30
28
Q
16
Q_N
W_{QN}
24
c_{QN}

Fig. 1B

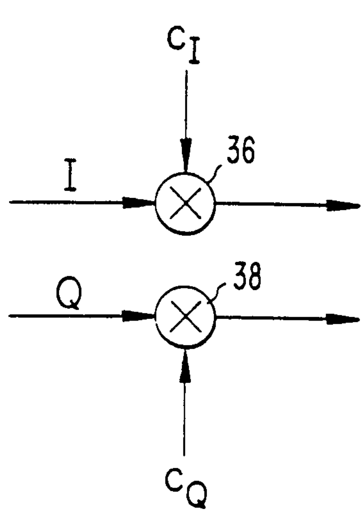

c_I
36
I
38
Q
c_Q

Fig. 2

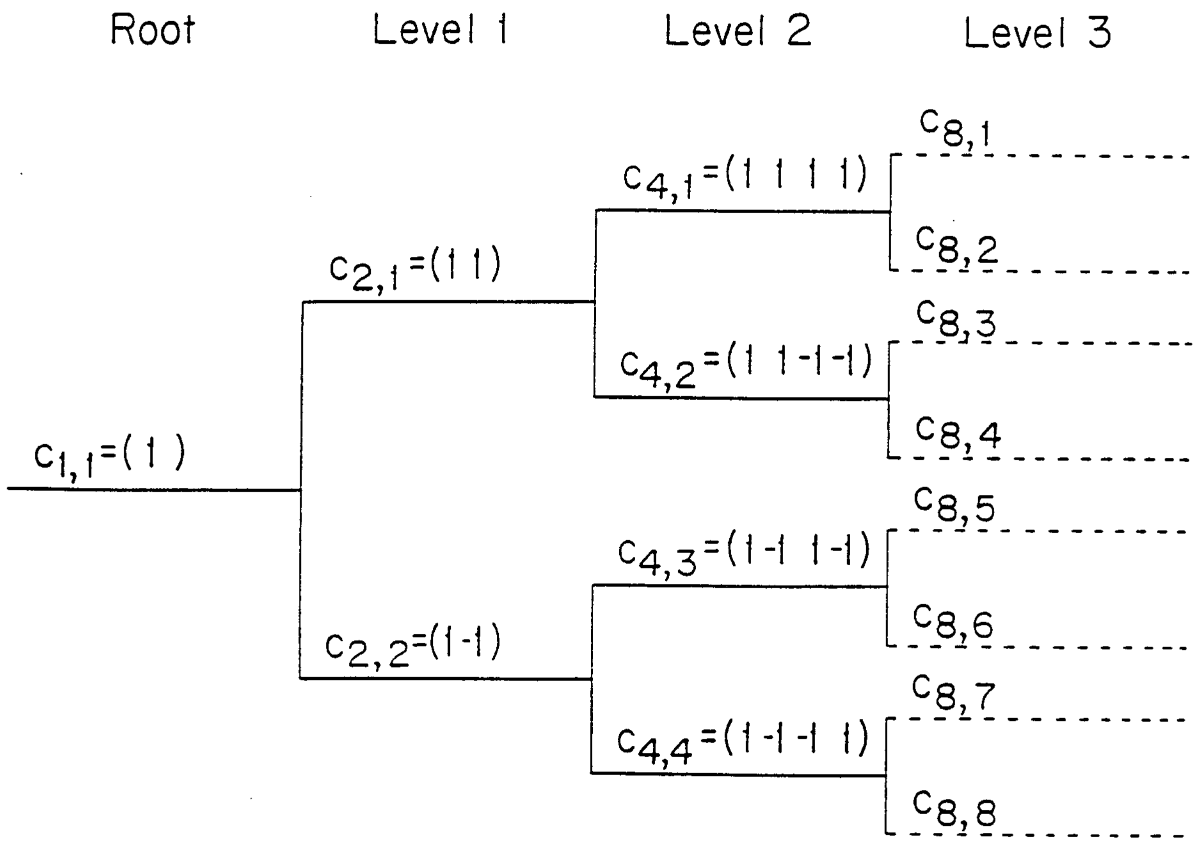
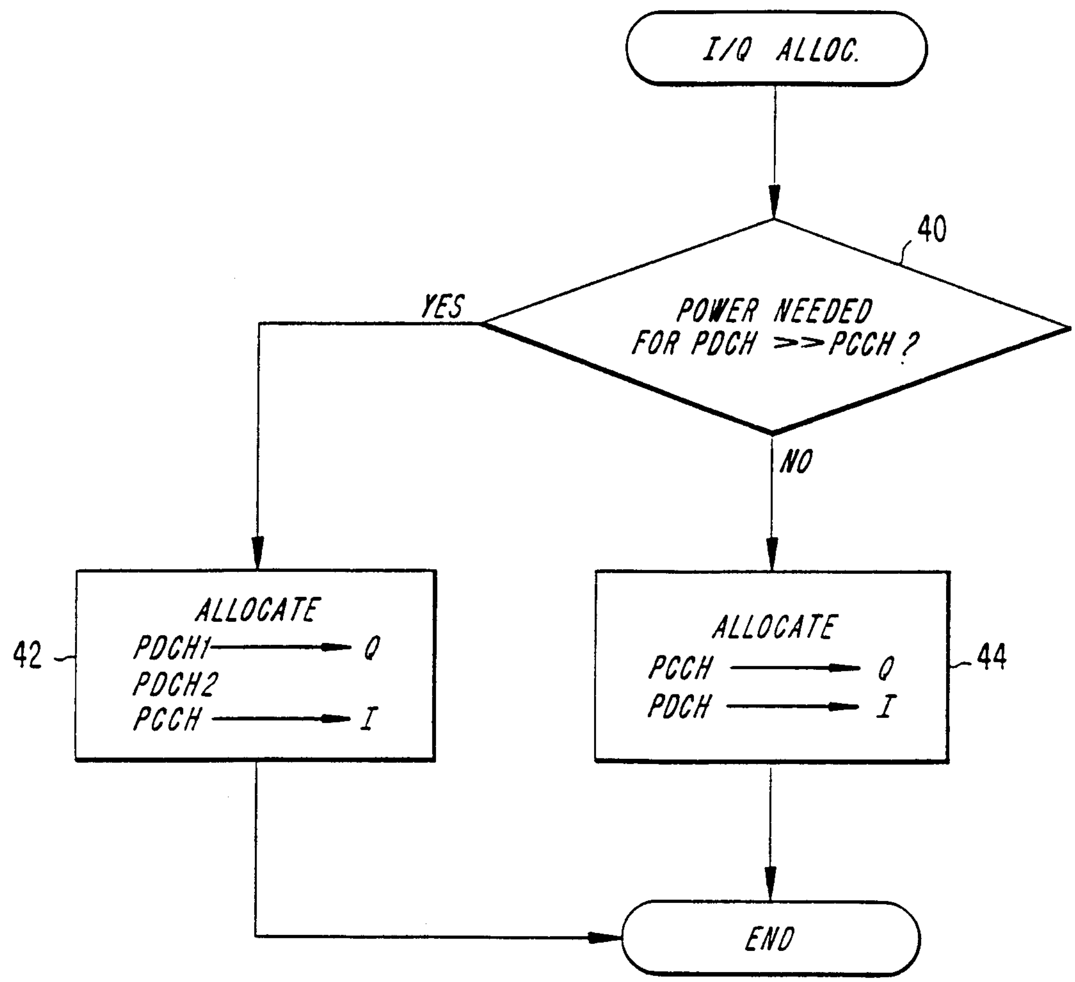
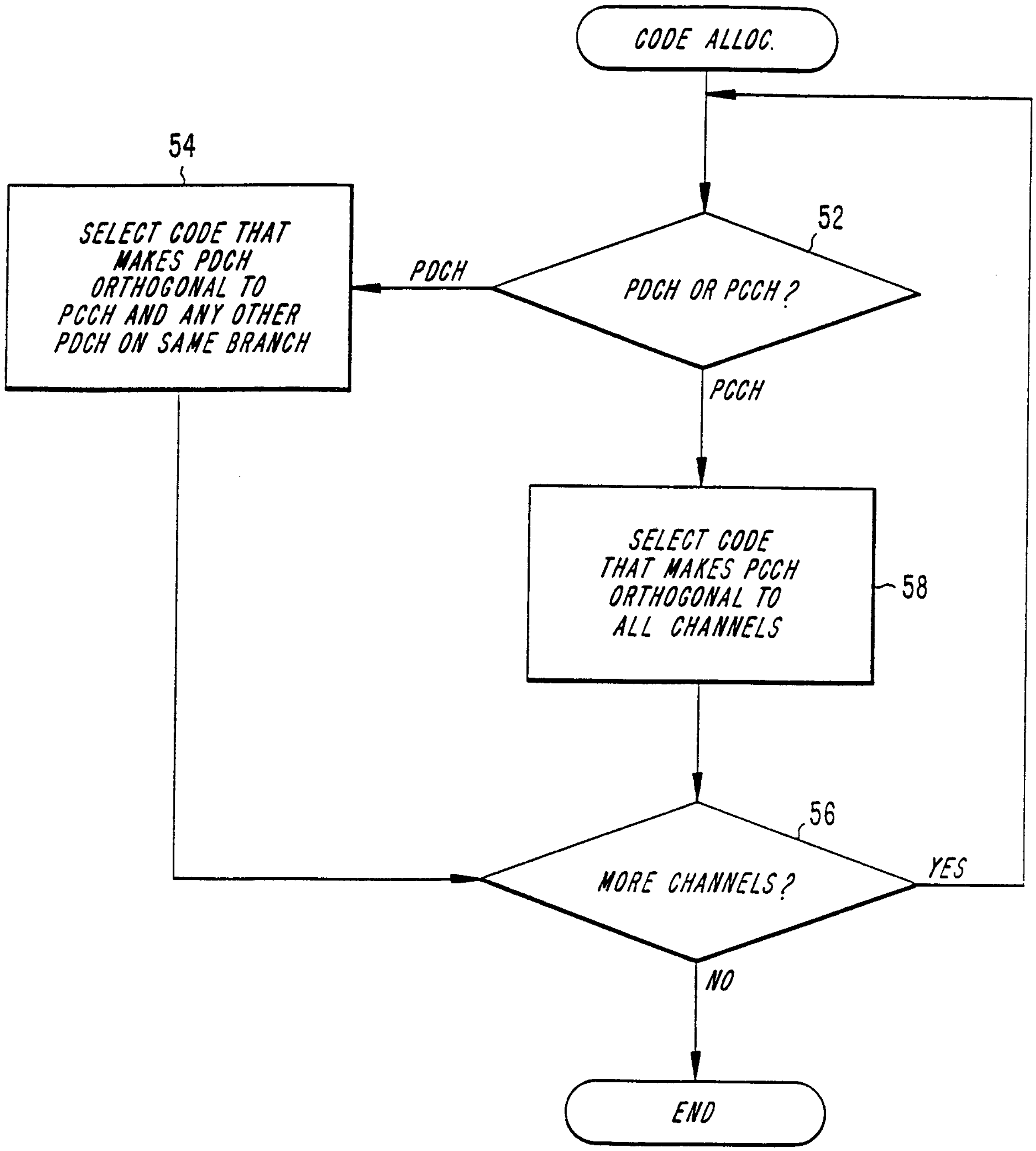

Root
Level 1
Level 2
Level 3
$c_{1,1}=(1)$
$c_{2,1}=(1\ 1)$
$c_{2,2}=(1\ -1)$
$c_{4,1}=(1\ 1\ 1\ 1)$
$c_{4,2}=(1\ 1\ -1\ -1)$
$c_{4,3}=(1\ -1\ 1\ -1)$
$c_{4,4}=(1\ -1\ -1\ 1)$
$c_{8,1}$
$c_{8,2}$
$c_{8,3}$
$c_{8,4}$
$c_{8,5}$
$c_{8,6}$
$c_{8,7}$
$c_{8,8}$

Fig. 3


I/Q ALLOC.
40
POWER NEEDED FOR PDCH >>PCCH ?
YES
NO
42
ALLOCATE
PDCH1 → Q
PDCH2
PCCH → I
ALLOCATE
PCCH → Q
PDCH → I
44
END



CODE ALLOC.
54
SELECT CODE THAT MAKES PDCH ORTHOGONAL TO PCCH AND ANY OTHER PDCH ON SAME BRANCH
PDCH
52
PDCH OR PCCH?
PCCH
SELECT CODE THAT MAKES PCCH ORTHOGONAL TO ALL CHANNELS
58
56
MORE CHANNELS?
YES
NO
END

Fig. 4

INTERNATIONAL SEARCH REPORT

Int ional Application No
PCT/SE 98/01317

A. CLASSIFICATION OF SUBJECT MATTER
IPC 6 H04J13/04 H04B7/26

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 6 H04J H04B

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US 5 559 788 A (ZSCHEILE JR JOHN W ET AL) 24 September 1996 see abstract see column 2, line 8 - line 20 see column 3, line 53 - column 4, line 12; figure 2 see column 5, line 33 - line 50; figures 6-8 ---	1,8,16
A	WO 95 03652 A (QUALCOMM INC) 2 February 1995 see abstract see page 4, line 3 - line 18 see page 14, line 32 - page 16, line 32; figure 2; table I --- -/--	1,8,16

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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Date of the actual completion of the international search	Date of mailing of the international search report
21 October 1998	26/10/1998

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL - 2280 HV Rijswijk Tel. (+31-70) 340-2040, Tx. 31 651 epo nl, Fax: (+31-70) 340-3016	Bossen, M

1

INTERNATIONAL SEARCH REPORT

International Application No
PCT/SE 98/01317

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
A	US 5 544 156 A (TEDER PAUL M ET AL) 6 August 1996 see column 3, line 49 - column 4, line 19 -----	2,4,8,16

1

Form PCT/ISA/210 (continuation of second sheet) (July 1992)

INTERNATIONAL SEARCH REPORT

Information on patent family members

Inte ,onal Application No
PCT/SE 98/01317

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WO 9503652	A	02-02-1995	AU	7368294	A	20-02-1995
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